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HIGH SENSITIVITY DC AMPLIFIER FOR BIOLOGICAL MEASUREMENTS

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PREPARED UNDER U. S. NAVY CONTRACT N6onr-264, T. O. 10

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Introduction

To improve the equipment available for zeta-potential studies at the Cornell University Medical College, a high sensitivity d-c amplifier was designed and constructed at the School of Electrical Engineering at Cornell. Designed specifically for the zeta-potential measurements the amplifier chopped the d-c signal, amplified it through sharply tuned twin-T amplifiers, and then synchronously converted it to d-c for meter measurement. The amplifier possessed a relatively high input impedance, a fractional microvolt sensitivity, and a band-width of less than one cycle. Though admirably suited to the measurement of steady state d-c microvolt signals this amplifier is not very adaptable for a large variety of biological measurements. The narrow bandwidth characteristic of the chopper-type amplifier prevents the observation of transient phenomena, and the slow response, even at hundred microvolt sensitivities, is unsatisfactory when many rapid readings of d-c potentials must be taken.

Considering the specialized applications and the restrictive characteristics of the chopper amplifier, the Medical College, under Naval Contract N6onr-264, T.O. 10, sought a complimentary equipment which would be adaptable for a larger variety of biological measurements. The amplifier described in this report was designed to fulfill this requirement of general applicability to biological investigation.

When the project was undertaken a specific completion date for delivery of a reliable engineered unit was specified. Although this completion date had to be postponed, the existence of such a specified amount of time influenced the design and construction of the amplifier to a great extent.

With an engineeringly sound piece of equipment as the ultimate goal of the project, the amount of circuit technique research was held to a minimum. It was early decided that the procedure most likely to produce a successful amplifier was the application of already developed circuit techniques. To this end an extensive search of the literature was made, and an abstracted bibliography, previously published*, was prepared by the author. From the literature studied the amplifier requirements were established, and the general circuitry was outlined.

General Performance Summary

The amplifier provides for measurements in the range from 10 microvolts to 100 millivolts with five sensitivity positions. Either single-ended or balanced input is available with an input resistance, limited by grid to ground resistors, of 10 megohms single-ended or 20 megohms balanced. The input capacitance between input cable lead and ground is less than 1.5 micromicrofarads. The net current flowing through the input grid resistors was 5×10^{-10} am-

*Interim Report, U. S. Navy Contract N6onr-264,T.O. 10, Cornell University, 1 Feb. 1951.

pere.

When the balanced input is used, the amplifier provides a high rejection of signals appearing in-phase upon both grids. Below 1,000 cycles per second an interfering signal below approximately one volt will give an output indication less than 1/20,000 of that given by an anti-phase signal of the same magnitude. Between 1,000 and 10,000 cycles per second the output indication of a common in-phase signal is not greater than 1/12,000 of that of an equal anti-phase signal.

The maximum response obtainable from the amplifier is down 3 db at between 40 and 60 kilocycles depending upon the sensitivity setting. As required by the phenomena under observation the 3 db attenuation frequently can be decreased to 10, 5, 1, and 0.5 kilocycles. The phase shift over the region from d-c to 30 kilocycles per second is approximately linear with frequency and ranges from 10 to 23 degrees at 10 kilocycles.

At the maximum sensitivity of 50 microvolts for full designed output a peak-to-peak drift of 100 microvolts was measured during a one-half hour period; the short time drift rate is, however, much higher. At a 250 microvolt sensitivity a peak-to-peak drift of 250 microvolts over a one hour period was recorded. A drift of less than 1.5 millivolts in 2.5 hours was measured at a sensitivity of one millivolt.

The amplifier has been designed primarily for use with auxiliary measuring or recording equipment having internal amplifiers;

consequently, the output voltages available at two independent outputs are relatively low and have a low source impedance. A balanced or unbalanced to ground output at an average ground potential is provided for oscillographic observation or measurement; since the designed unbalanced output voltage is only one volt an oscilloscope employing a d-c deflection amplifier must be used. The same input signal provides an approximately four volt output at an average of +100 volts above ground for meter deflection measurement.

Requirements

Bandwidth

For such steady state d-c measurements as those of zeta or streaming potentials a response from d-c to one or two cycles is adequate. As noted previously, however, this restrictive bandwidth with the inherent slow response, while adequate, is undesirable.

General diagnostic techniques such as electrocardiography and electroencephalography are concerned with a relatively narrow frequency range from about one to 50 or 60 cycles per second.[6.17*] Although a capacitance coupled amplifier can be designed for this work the low frequency response must be extended much below one cycle. It has been pointed out by Dawson[6.33] that both electrocardiograph and electroencephalegraph recordings are seriously modified by phase shift. Consequently, low frequency responses in the order of 0.1 cycle have been employed for EEG amplifiers[1.17].

Electroneurophysiological research in action potential measure-

*Refers to bibliography previously cited. 6.17 means category No. 6 and the No. 17 abstract in this category. All references not otherwise identified refer to the same bibliography.

ment demands a relatively wide response for investigation of transient phenomena. A typical action potential recording consists of a rapidly rising and falling "spike" which drops below the resting potential by 0.1 to 0.4% of the spike crest[7.3] and then gradually returns to the resting potential; this latter portion is called the "after potential." An action potential spike can attain its crest in 0.2 to 0.1 millisecond[7.1] and last approximately 0.4 millisecond [7.2, 7.3]. The after potential (rabbit cervical sympathetic nerve) lasts about 80 milliseconds[7.2]. To record faithfully such bioelectric potentials an amplifier response flat from d-c to 10 kilocycles per second has been considered necessary[1.3, 1.23, 6.15, 6.17] although as late as 1942 an upper limit of 8 k.c. was considered unnecessarily high[6.23]. A recent amplifier described by Grundfest[6.1] is flat to 30 kilocycles. A rise time of less than 10 microseconds has been deemed necessary[6.1].

For action potential measurements sufficient low frequency response can be obtained with capacitance-coupled amplifiers. Investigative techniques, however, require that there be no blocking with 20 to 50 volt stimulating signals and that the amplifier recover immediately from such signals[6.1] since the action potential may follow the stimulating voltage within 0.1 millisecond[1.23]. This requirement precludes the use of capacitance coupling and dictates a direct coupled amplifier.

Input Impedance

Since the subject plus the electrodes has inherently poor regulation[6.15] and has a resistance of the order of megohms,[6.1] a biological amplifier requires a high input impedance. Bishop and Harris[1.3] specify an input impedance requirement of greater than 10 megohms in the range 0 to 10 kilocycles; a recent amplifier described by them has an input impedance of 100 megohms at 10 kilocycles. Grundfest[1.6] employs 10 megohm resistors in a recent amplifier to give a balanced input resistance of 20 megohms between electrodes.

Sensitivity

Representative measurements in electrocardiography are in the region of one or two millivolts, while electroencephalographic potentials range from 10 microvolts to 1 millivolt[6.17]. Gasser and Grundfest[7.3] mention action potentials of 0.1 and 12 millivolts from Mammalian A fibers and Parnum[6.15] gives 1 millivolt as a typical peak value for action potentials. A sensitivity requirement of the order of 50 microvolts to 100 millivolts at d-c has recently been stated by Grundfest[6.1]. Amplifiers for biological work have been described with maximum sensitivities of 15 microvolts for full scale output,[1.17] 25 microvolts per inch[1.6], and 10 microvolts for full scale oscilloscope deflection.[1.8]

Stability

A useful stability limit for a direct coupled biological amplifier is set at a drift of the order of 100 microvolts peak-to-peak over 30 minutes[1.3]. Since fluctuations in contact potential of tissue and electrodes are of the order of 50 microvolts and are unpredictable in their course, a short term stability within 50 microvolts for one minute is specified by Grundfest[6.1]. In a recent amplifier[1.6] Grundfest achieved an average drift of 40 microvolts per minute with fluctuations of 800 microvolts over 1/2 minute.

Discrimination

One of the troublesome factors in bio-electric measurement is the difficulty of shielding the subject under investigation. As a consequence it is frequently necessary to observe small signals in the presence of large pick-up voltages. These interfering voltages may be of the order of hundredths of tenths of a volt and are approximately in phase[6.21, 6.22]. Parnum points out that the electrode pick-up is several millivolts and is not sinusoidal[6.15]. The impossibility of using a common ground path between amplifiers was early demonstrated by Adrian and Matthews[8.0]. Three pairs of electrodes (one of each pair connected to a common ground) were placed one on the heart and two on the liver; all three showed the potential waves of the electrocardiagram. The use of a differential amplifier eliminated the effect of potential changing of the grounded lead.

Dawson and Walter[1.20] state that high discrimination ratios (ratio of differential gain to in-phase gain) of 1,000 to 1 or greater can rarely be maintained in practice owing to uncontrollable variations in electrode characteristics. They indicate that such errors are of the order of 1 part in 1,000 depending upon the type of electrode and the input impedance. Goldberg reported[1.23] a rejection ratio of 100,000 to 1 when his amplifier was properly balanced, and Grundfest achieved a differential out-of-phase to common mode amplification of greater than 100,000 at d-c which dropped to about 10,000 at high frequencies[1.6].

Grid Current

It has been pointed out by Grundfest[1.6, 6.1] that grid current flowing in the external circuit must be less than 10^{-10} ampere; currents in the order of 10^{-8} to 10^{-9} ampere often change the characteristics of tissue.

Miscellaneous

As practical considerations in the design of an amplifier for brain potential measurement, Bishop and Harris[1.3] indicate the following: a) sensitivity to external disturbances should be a minimum without bulky screening, b) the output level should not be "excessive" with respect to ground, c) the setting-up procedure should be simple, and d) the amplifier should be power supply operated.

Design Specifications

After a review of the requirements demanded of an amplifier for general biological measurement the following specifications were established.

a) Sensitivity: 10 microvolts to give full scale output with a maximum input of 100 millivolts.

b) Input impedance: 10 megohms in the range 0 to 10 kilocycles.

c) Frequency response: flat from 0 to 10 kilocycles.

d) Input grid current: 10^{-10} ampere.

e) Drift: less than 200 microvolts peak-to-peak over 30 minutes, and not over 50 microvolts per minute.

f) Rise time: 10 microseconds.

g) In-phase to out-of-phase rejection ratio: 3,000.

h) Power supply: completely line operated.

CIRCUIT DESCRIPTION

The Amplifier

The amplifier consists entirely of push-pull stages from input to output circuits. The advantages gained from the use of push-pull stages include the following:

a) Stability against variations in plate supply voltage and some balancing out of the effects of heater voltage fluctuation. Gray[5.8] points out that the use of two tubes multiplies the probable short time drift by $\sqrt{2}$ but that for longer periods there is some cancellation between double triodes.

b) Permits the use of large common cathode resistors which give a high ratio of anti-phase to in-phase amplification.

c) Makes possible the use of differential gain control schemes which change amplification but do not affect the average d-c levels.

d) Simplifies the supplying of pentode screen grids with a constant voltage.

e) Maintains current drawn from power supply constant.

An input cathode follower is followed by two triode stages of amplification and then by two pentode amplifying stages. The output of the amplifier proper is then fed to an output circuit chassis where cathode followers provide low impedance outputs for meter deflection or cathode ray oscillograph measurement. A detailed description of the amplifier circuit (Fig. 1) follows.

Input Circuit

To meet the input impedance requirement of 10 megohms over the range 0 to 10 kilocycles an input resistance at d-c considerably higher than 10 megohms is needed to allow for capacitance shunting. Since grounded input grid resistors were desired 10 megohm grid to ground resistors were used to give a balanced signal input resistance of 20 megohms. To maintain an input impedance of 10 megohms at 10 kilocycles this permits a shunting capacitance of 1.5 micromicrofarads. Since it was desired to maintain the 10 megohms with approximately $2\frac{1}{2}$ feet of shielded cable to be provided, the double shielded driven shield technique described by Daniels[6.19] was used. By using a cathode follower with the inner shield of the double shield input cable connected to the cathode, the input capacitance can be reduced by the factor, a , given by

$$a = 1 - \frac{1}{1 + 1/\mu + \frac{1}{gmZ_L}}$$

where Z_L is the impedance of the cathode resistor and shunting capacitance. For the 12SL7 used the value of a is approximately 0.02 which sufficiently reduces the input cable capacitance of 40 to 50 micromicrofarads. A duo-triode is used since the two triode sections must operate with biases sufficiently alike that the succeed-

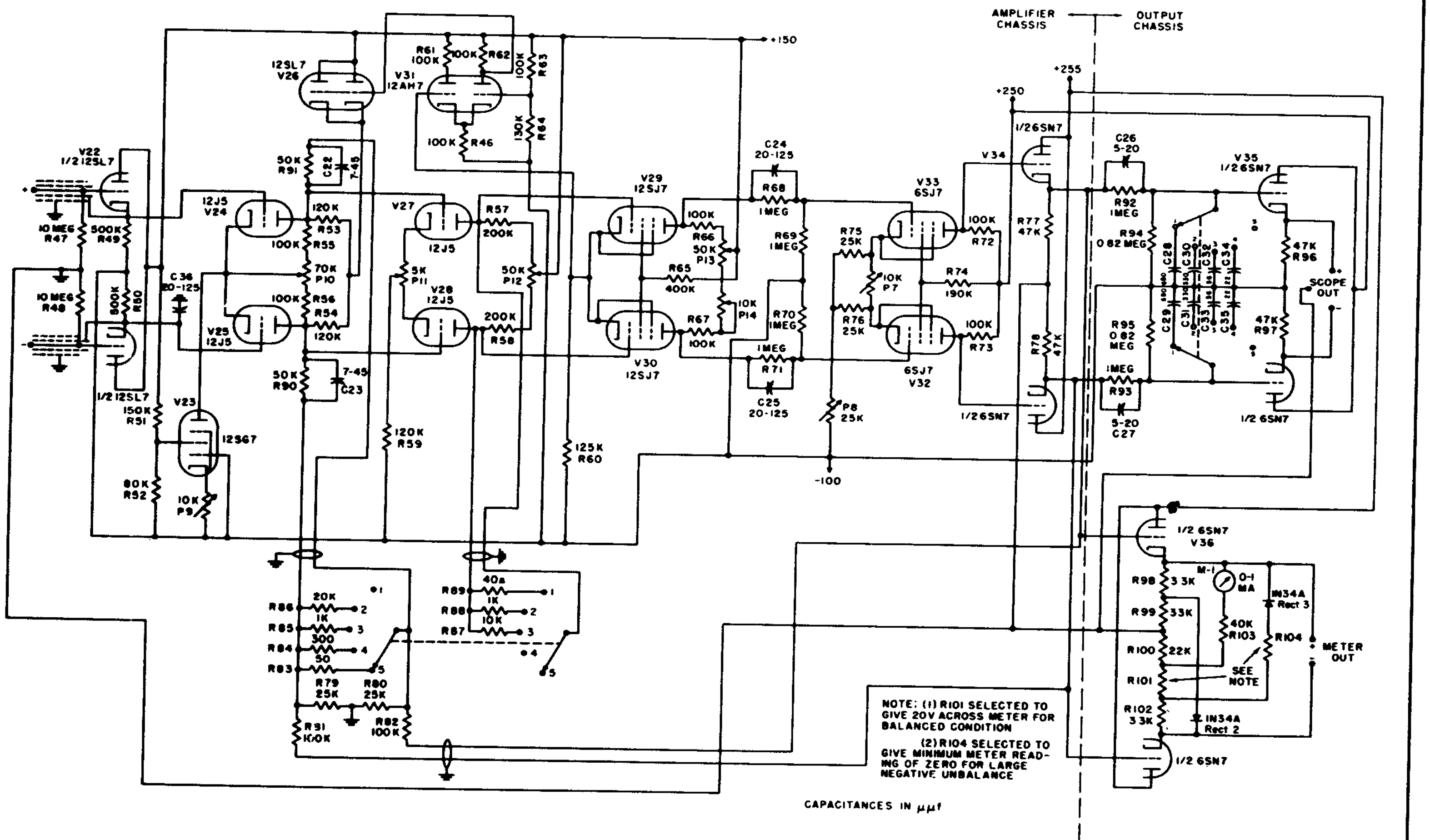


FIG.-1 DC AMPLIFIER

ing stages will not be unbalanced beyond the range of balance controls.

One of the most important characteristics of a d-c amplifier for biological measurement is its ability to discriminate against signals appearing in-phase on both input grids. This is accomplished by the use of cathode coupled 12J5's, V-24 and V-25, in a circuit described by Parnum[1.5].

By the use of a sufficiently large cathode resistor the in-phase amplification of a cathode coupled stage can be reduced to a value depending primarily upon the magnitude of the negative voltage supply. However, as Parnum points out, unequal circuit values in the two halves of a differential amplifier stage give rise to an out-of-phase voltage even though the in-phase amplification is zero. As a figure of merit Parnum defines the term "Transmission Factor," H, given by H = M/K where M is the out-of-phase gain and K is the ratio of the in-phase input signal to the resultant out-of-phase output signal. The transmission factor should thus be as high as obtainable. H for a triode differential amplifier stage is shown [1.5] to be given by

$$H = \frac{r_p + R + ZR_c(\mu+1)}{r_p} \Big/ \left[\frac{\Delta R}{R} - \frac{\Delta r_p}{r_p} + \frac{\Delta \mu}{\mu} \left(\frac{r_p + R + ZR_c}{r_p} \right) \right]$$

where R is the plate load resistor and R_c is the common cathode resistor. To make the denominator as small as possible it is evident that $\frac{\Delta\mu}{\mu}$ should be made small because of its large multiplying factor. A large cathode resistor also increases the numerator because of its large multiplying factor $2(\mu+1)$.

In the amplifier circuit resistors R-55 and R-54 together with potentiometer P-10 shunt tubes V-24 and V-25 and give a reduced effective μ which is varied by P-10. The new effective μ of the tube is given by $\mu' = \mu \frac{S}{S + rp}$ where S is the tube shunting resistance, rp is the tube plate resistance, and μ is the amplification factor of the tube. Since the adjustment of P-10 is fairly critical the 70k ohms selected permits adjustment of μ for tubes with μ differing by 10%. Since normal variation of μ is $\pm$ 20% this entails the selection of 12J5's for this stage which have μ's differing by less than 10%. The reduced μ' of 18 gives a first stage gain of 16.

Since the four input cables were constructed from single shielded cable and an additional shield, it was found that the capacitance between the inner shield and the outer shield of any pair of cables differed by not more than approximately 100 micromicrofarads. This unequal shunting capacitance across the input cathode follower load resistors introduced an unbalanced signal to the grids of the first stage tubes, and this, in turn, lowered the transmission factor. To obtain a capacitance balance the variable ceramic capacitor, C-36, was added between one cathode of V-22 and ground. The input cables were then selected so that the one with the lesser capacitance was installed on the input terminal supplying the signal to the cathode of V-22 with the capacitor.

For a desired over-all H of 50,000 Parnum demonstrates[1.5] that the least troublesome solution is a pentode cathode load and shunt control for the first stage with a large cathode resistor for the second stage. It is essential that the in-phase amplification of the first stage be held to a minimum since the value of K for an unadjusted second stage will be relatively small, and the out-of-phase output can be kept small only by keeping the in-phase signal to the second stage small. The use of the 12SG7 pentode as a cathode load permits a reasonable current flow through the relatively low static resistance of the tube but provides a high dynamic resistance of approximately 3 megohms.

<u>The Second and Third Stages</u>

The second stage of amplification, with a gain of 19, is another 12J5 stage with a 5K potentiometer, P-11, between the cathodes and a large common cathode resistor. The P-11 adjustment is a chassis control used for initial balancing to compensate for the unbalance from the input 12SL7 and the d-c unbalance introduced by P-10 in the adjustment of the amplification factor.

The main amplifier balance is provided by a 10 turn potentiometer, P-12, placed in the plate load circuit of the second stage tubes, V-27 and V-28.

Both input and second stages employ 12J5 medium triodes which have the following advantages: 1) triodes introduce less tube noise than the multi-grid pentodes, 2) the construction of a medium μ triode is less susceptible to microphonics. 3) a medium μ tube will have smaller variations than a higher μ tube and will thus simplify the selection of sufficiently balanced tubes and, 4) the use of individual rather than dual triodes permits more flexibility in matching tubes for μ and/or for static d-c currents.

The third stage is a 12SJ7 pentode amplifier with a large common cathode resistor and with two adjustments in the plate load circuit. The 50K potentiometer, P-13, is a chassis adjustment to balance the third and fourth stages and the 10K potentiometer, P-14, is a panel adjustment for fine control of the output circuit balance. The gain of the third stage is 120.

To provide stabilization of the mean output level of the second stage a feedback circuit is used which was described by Bishop and Harris[1.15] and employed in an amplifier of their design[1.3]. However, instead of supplying the plates of the first stage directly from the 12SL7 cathode follower, V-26, the 12AH7 dual triode provides an amplified voltage to the cathode follower. Although the 12AH7, V-31, amplifies the voltage changes at the cathodes of the third stage by a factor of 15 its primary function is to provide a higher voltage of the same polarity to the grids of V-26 than is obtainable at the cathodes of V-29 and V-30 directly.

By supplying the plate power to the first stage with the cathode follower arrangement the first three stages can be fed from the +150 volt regulator without coupling troubles through the power supply. The plate currents for the first three stages are held in the range 0.1 to 1.0 milliamperes since Gray[5.8] points out that drift studies indicate that for most ordinary receiving tubes the minimum drift occurs at plate currents between those values and at plate and screen voltages as low as possible.

For stability the tubes of the first three amplifying stages, the amplifier and cathode follower of the first stage supply, the input cathode follower, and the pentode cathode load are all heated from the regulated 38 volt d-c supply. The heaters are connected in series and parallel arrangements with the heaters of V-24 and V-25 in series and the heaters of V-27 and V-28 in series so that the same current flows through the heaters of both tubes of a single stage.

Final Amplifier Stage

The output of the third stage is coupled to the grids of the final amplifiers, 6SJ7's V-32 and V-33, through the voltage dividers R-68 to R-71. Adjustable ceramic capacitors C-24 and C-25 are used to compensate for the capacitance shunting effect of the tubes upon resistors R-69 and R-70.

The final stage is operated under conditions specified by Vacuum Tube Handbook resistance coupled amplifier data to give a gain of 110 with the cathodes tied together. Two adjustments are provided on the final amplifier stage: an adjustable common cathode resistance, P-8, and a differential type adjustable gain control. Although this type of gain control gives no improvement in discriminating ratio with decrease in gain, it has the advantage of not disturbing the d-c levels as the gain is varied. The panel mounted control, P-7, provides a limited amount of gain adjustment as described later. The adjustable chassis mounted control, P-8, varies the plate voltages of the stage and this ultimately adjusts the final output voltage level of the output stages.

Cathode Follower Stage

The output of the 6SJ7's is coupled directly to the grids of a duo-triode 6SN7, V-34. This provides a low impedance output to a separate output circuit chassis and provides the feedback signal. This final tube receives its plate voltage from the unregulated +255 volt supply which has the negative side grounded. The output of the cathode follower is at an average potential of 106 volts above ground.

Gain Control

The gain of the amplifier from balanced input to balanced output of the amplifier cathode follower is approximately 2×10^6. Experience demonstrated that a full scale sensitivity of 10 microvolts as originally proposed was not useable. A maximum sensitivity of 50 microvolts for full output of 2 volts was deemed the maximum practicable. The 2 volts balanced, or 1 volt unbalanced, output was adequate for use with an oscilloscope providing a self-contained d-c amplifier. Since the output circuits have a gain of 0.45 the maximum amplifier gain desired is, therefore, 8.9×10^4.

To obtain the gain reduction negative feedback is used from the output of the cathode follower, V-34, to the grids of the second stage amplifiers, V-27 and V-28. The feedback is not carried to the input of the first stage because it would upset the μ balance of the first stage and decrease the transmission factor. To introduce the feedback signal to the input the feedback paths would have to be balanced carefully, and no adjustments could have been included inside of feedback portion of the amplifier.

Bishop and Harris[1.3] point out that the feedback should be degenerative for mean d-c levels as well as for differential changes. The mean d-c level for the feedback signal is determined by resistors R-79 to R-82 which form a voltage divider across the V-34 output to give + 21 volts at the point from which the feedback signal is taken to match the 21 volt grid voltage of V-27 and V-28. Feedback for mean d-c level is thus a constant portion of the output signal. To get the approximately 1.5×10^{-4} feedback factor required differentially, resistors R-79 and R-80 are shunted by R-83, a 50 ohm resistor whose final value was determined experimentally. The value of R-83 was determined with the fine gain control, P-7, located in its mid-position.

Five values of gain were desired corresponding to 2 volts output for 50 and 250 microvolts and for 1, 10, and 100 millivolts. By increasing the shunting resistor to 300 ohms the feedback can be increased sufficiently to obtain the 250 microvolt sensitivity with no further amplifier change. However, increasing the feedback sufficiently to obtain a 1 millivolt sensitivity resulted in strong amplifier oscillation in the vicinity of 200 to 300 kilocycles. For the three less sensitive gain values it was necessary to decrease the amplification of the second stage as well as to increase the feedback factor. The second stage gain was decreased by the use of resistors shunting the plate load resistors. Although this method has the disadvantage of decreasing the differential gain without affecting a corresponding decrease of the in-phase gain, it has the virtue of not affecting the average d-c levels of the amplifier. The in-phase amplification is so low that the shunt type of gain reduction did not adversely affect the amplifier's performance.

To provide for gain variation resulting from resistor changes or tube aging or replacement and to permit calibration of the amplifier at other than the nominal sensitivities the fine gain control of P-7 was provided. Since it is included inside the feedback portion of the amplifier its range of effectiveness is dependent upon the sensitivity setting.

The range of sensitivity provided by the Fine Gain Control was measured to be the following:

Amplifier Sensitivity Setting	Fine Gain Control Maximum	Fine Gain Control Minimum
50 μv.	100 μv.	45 μv.
250 μv.	265 μv.	215 μv.
1 mv.	1.2 mv.	0.73 mv.
10 mv.	12.8 mv.	8.8 mv.
100 mv.	122 mv.	33 mv.

At the lowest gain setting the second stage is reduced so much that the main balance control, P-12, in the plate circuit of that stage becomes ineffective. It was necessary to provide an additional panel control, P-14, to balance the output of the third and fourth stages separately.

To correct a tendency toward instability on the 250 microvolt sensitivity position, variable ceramic capacitors C-22 and C-23 were shunted across R-90 and R-91 in the feedback circuit. During the adjustment of the amplifier the two capacitors are increased equally from their minimum value to remove any peak in the response in the vicinity of 100 to 300 kilocycles with the amplifier set at the 250 microvolt sensitivity.

Output Circuits

The signal from the cathode follower, V-34, is fed to a separate chassis containing the output circuits and the frequency response limiting circuits. The voltage dividers consisting of R-92 and R-94, R-93 and R-95 drop the d-c level of the signal from +106 to approximately -7 volts. This provides the correct bias to place the cathodes of V-35 at an average ground potential. This drop in d-c level is obtained, however, at the expense of a gain of 0.45. The output of V-35 provides either a balanced signal of 2 volts or a 1 volt signal of either polarity with respect to ground.

Capacitors C-26 and C-27 compensate for the tube and switch capacitances shunting the resistors R-94 and R-95 and are adjusted by minimizing the phase shift through the output circuit alone. Capacitors C-28 through C-35 provide five bandwidth positions. One gives the maximum response available from the amplifier while the other four seccessively decrease the 3 db response frequencies to

10, 5, 1, and 0.5 kilocycles.

The same signal from V-34 which is fed to the voltage divider mentioned above is also fed directly to the grids of the push-pull cathode follower V-36. The cathode load resistors are grounded and the plate is supplied from the unregulated 255 volt power supply. Output terminals from each cathode are provided for meter deflection measurements; this output is at an average potential of 108 volts above ground.

A meter, M-1, is provided as a panel indicator for balancing the amplifier and as a protection before connecting an external meter to the output terminals. Although a large unbalance is possible a reasonably sensitive balance indicator was desired. Since no zero center scale meter was available a standard 0 - 1 milliampere meter was used with a 40,000 ohm resistor, R-103, and was connected so that it read half-scale when the cathodes were at the same potential. Two germanium diodes, Rect-2 and Rect-3, are connected with a bias of 10 volts so full meter sensitivity is maintained for unbalances of + or - 10 volts. Beyond that the diode circuits shunt the meter. R-104 in series with Rect-3 acts as one arm of a bridge circuit which limits the minimum current through the meter to zero - the needle does not go off scale. For large unbalance in the positive direction, however, the diode, Rect-2, provides a shunting path around the meter and limits the meter current to approximately 1.5 milliamperes at maximum unbalance encountered.

Power Supplies

The d-c amplifier was designed for use in a location with poor voltage regulation and only moderate frequency control. Since such a location dictates the use of a-c voltage regulating equipment for precise measurements with electronic apparatus, it was decided to depend upon such separate a-c regulators for the first stage of voltage regulation. Since both voltage and frequency varied it was planned that the d-c amplifier would be supplied a-c power from a line voltage regulator of the electronic rather than resonant circuit type.

In both heater and plate power supplies each shunt amplifier and regulator stage employs two identical tubes which are cathode coupled. Although this has the disadvantage of doubling the number of tubes and thus doubling the regulated heater current required, the use of cathode coupling provides increased regulator stability, and, by the use of proper cathode resistor, the voltage of the cathode can be adjusted to the necessary circuit value with no loss in stage amplification. Also, the push-pull nature of the circuit maintains the current drain on the power supply constant and thus helps to improve the stability of the output voltage of the power supply.

The power supply circuits - both heater and plate - are based on a power supply for an electroencephalograph described by Johnston 1.17.

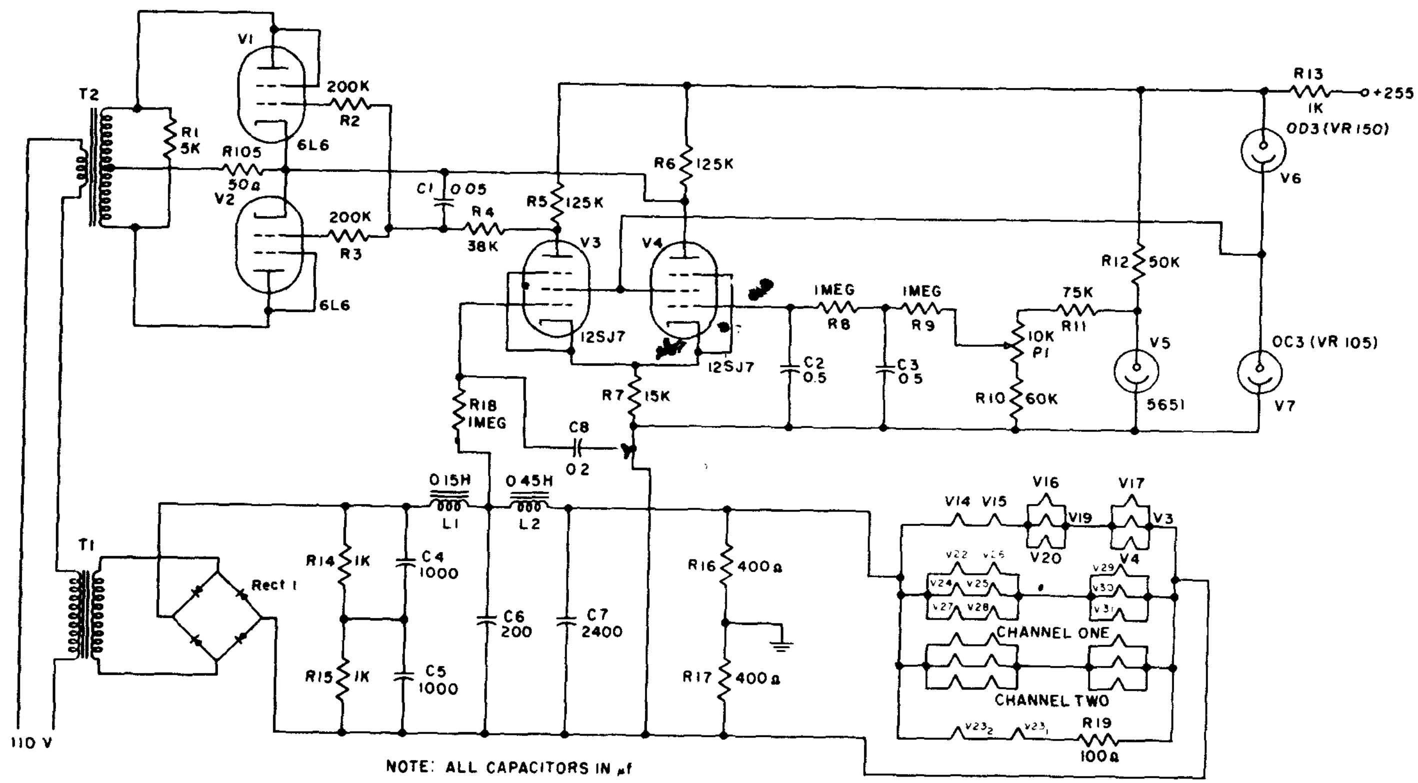
Heater Power Supply

In a high gain d-c amplifier it is essential that the initial amplifying tubes and the critical power supply regulator amplifiers have a direct current heater supply. This is necessary to prevent the introduction of power supply interference from heater lead pick-up or from vacuum tube inter-electrode leakage.

To provide the direct current for the critical tube heaters a series transformer regulator patterned after the one described by Johnston[1.17] is used.

Johnston used a supply voltage of 220 volts obtained from a transformer built into the encephalographic apparatus proper. Although the use of 220 volts would have permitted the use of smaller currents in the regulating load, it was decided to operate the regulator directly from 110 volts and eliminate the need for an additional step-up power transformer.

The power supply (Fig. 2) provides 1.5 amperes at 38 volts d-c at +23 and -15 volts with respect to ground. Transformer T-1 with a ratio of approximately 1.3:1 operates with a primary voltage of 60 volts. Its output is rectified by a bridge connected selenium rectifier, Rect-1, and fed into a capacitor input filter with two smoothing sections. Equal voltage is maintained across C-4 and C-5 by resistors R-14 and R-15. Although better filtering is obtainable with a choke input filter the rectifier input voltage required exceeded the voltage capability of the selenium rectifier available. Subsequent use showed that with the capacitor input no power supply



NOTE: ALL CAPACITORS IN μf

FIG. 2 DC FILAMENT POWER SUPPLY

interference was present in the amplifier output. Resistors R-16 and R-17 reference the output to ground and provide the ground return for the regulating amplifiers.

Regulation of the output voltage is obtained through transformer T-2 which is connected across the supply line in series with the main transformer T-1. T-2 is a nominal 110 volts to 1200 volts center-tapped power transformer which is operated here with about 50 volts across the primary. The secondary is loaded by a 5,000 ohm resistor, R-1, directly across the secondary and by a 6L6 across each half of the secondary. Tubes V-3 and V-4 vary the bias on V-1 and V-2 to control the load current flowing the 6L6's in accordance with the output voltage of the filter, and thus the primary voltage across T-1 can be maintained constant. To keep from exceeding the permissible power dissipation of the 6L6's the additional loading required was obtained by the use of R-1 in preference to employing two other tubes in parallel with V-1 and V-2.

Any voltage change occurring at the center of the smoothing filter is amplified by the cathode coupled degenerative amplifiers V-3 and V-4. The signal is then coupled to V-1 and V-2 through the low pass filter consisting of R-4 and C-1 which prevents oscillation of the circuit at a high frequency. To stop a persistent low frequency oscillation at around 15 cycles an additional filter consisting of R-18 and C-8 was added.

The plate and screen power for the shunt amplifier is provided at + 255 volts from an unregulated supply with the negative side grounded. Regulation for tubes V-3 and V-4 is obtained by use of the series resistor R-13 and the voltage regulator tubes V-6 and V-7. The screens of the amplifier tubes are supplied from the junction of V-6 and V-7 to provide a constant screen voltage. An RCA 5651 voltage reference tube is operated from the regulated side of R-13 through R-12 and provides the reference voltage for the regulating amplifier through potentiometer P-1 and resistors R-11 and R-10. The reference voltage is fed to the grid of V-4 through a two section R-C low pass filter to reduce any rapid voltage fluctuations occurring in the 5651.

To improve the stability of the output voltage the cathodes of the degenerative amplifiers, V-3 and V-4, are heated from the regulated d-c output of the power supply.

<u>Plate and Negative Voltage Power Supply</u>

The plate and negative voltage supplies for the d-c amplifier had to fulfill the following requirements:

1) To permit the use of large common cathode resistors in the amplifier a negative voltage of -100 volts was needed.

2) Since the plate voltages in a direct coupled amplifier become successively higher a reasonably high value of regulated plate supply is necessary. A singly regulated supply at + 250 volts and a doubly regulated + 150 volts were required.

3) A single plate supply must feed two identical amplifier channels with no interaction between the amplifiers through the power supply.

4) Since ordinary decoupling methods are not applicable to direct coupled amplifiers, any one section of the plate supply can directly feed only two successive stages of the amplifier.

5) The regulator must operate from the output of a 110 volt a-c line regulator.

An electroencephalograph described by Johnston[1.17] had a low frequency response down to 0.15 cycle with a full scale sensitivity of 15 microvolts; the amplifier stages were direct coupled except for one capacitance coupled stage. It was felt that the power supply designed for this amplifier would satisfactorily serve as a model for the design of the plate and negative voltage power supply for the d-c amplifier under discussion.

Transformer T-3 (Fig. 3) provides a secondary voltage of 1000 volts center-tapped which is rectified by a 5U4 and smoothed by a two-section capacitor input filter. A 30 second thermal time delay relay, Re-1, inserted ahead of the last filter output capacitor delays application of the high voltage until the tubes are heated. Two 6L6's (V-9 and V-10) triode connected and operating in parallel serve as series regulating tubes with a voltage drop of 220 volts across them. They maintain the output of the first section of regulation at a potential difference of 350 volts with a current of 120 milliamperes.

Two cathode-coupled 6SJ7's, V-12 and V-13, act as degenerative amplifiers for the first section of the regulator. The reference

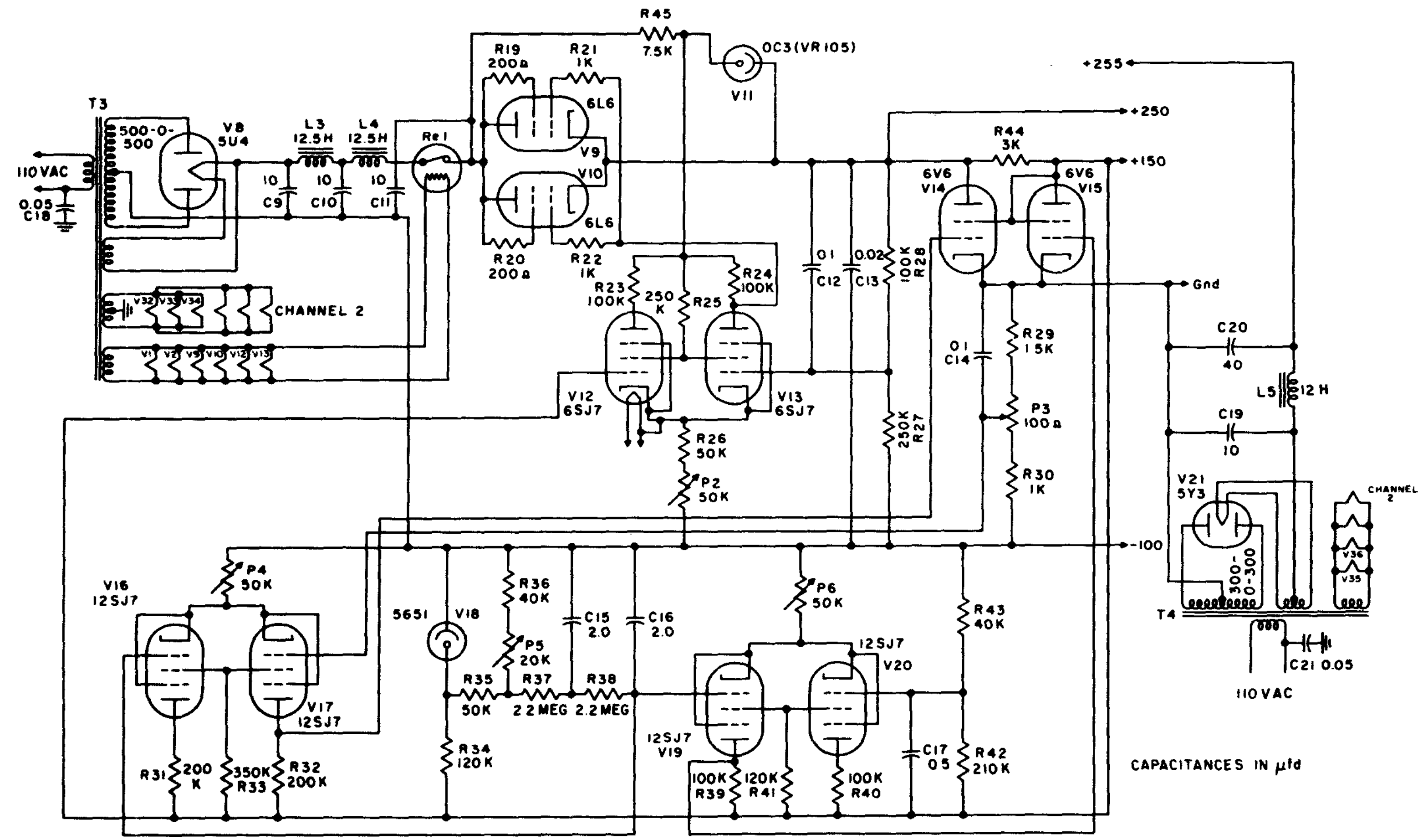


FIG-3 PLATE AND NEGATIVE VOLTAGE POWER SUPPLIES voltage for V-12 is the final, doubly regulated + 150 volts; the signal voltage for V-13 is taken from the voltage divider R-27 and R-28 across the output of the section. Capacitor C-13 decreases the high frequency impedance of the regulator, and C-12 shunts R-28 and increases the a-c signal on the grid of V-13.

The plates and screens of V-12 and V-13 are fed from the filter output by R-45 and the voltage regulator tube V-11. This gives a higher supply for the plates and screens than the regulator output with better regulation than obtained from feeding from the filter output. The 20 volts bias for the 6L6's can thus be obtained with the use of large plate load resistors and a resultant gain in amplification.

The output of the first regulator is 350 volts with neither side referenced to ground. The second section of the regulator uses two 6V6's, V-14 and V-15, to reference the negative side at a regulated -100 volts and to provide a doubly regulated + 150 volt supply. Since two series type regulators in tandem are prone to oscillation at a high audio frequency[1.17], the second stage of regulation is done by using shunt rather than series regulators.

The cathodes of both 6V6's are grounded through the ground wire of the amplifier. The shunting currents through V-14 and V-15 return to the negative side of the filter through R-29, P-3, and R-30. Potentiometer P-3 provides the regulating signal which is amplified by the 12SJ7's V-16 and V-17 and applied to the grid of V-14 to vary the current and thus maintain a drop of 100 volts across R-29, P-3, R-30.

The +150 volt output of the power supply is regulated at 250 ts above the -100 volt output by varying the shunting current wn by V-15 and thus varying the IR drop across R-44 to maintain output voltage constant. The regulating signal for controlling 5 is taken from the voltage divider R-42 and R-43, amplified by cathode coupled 12SJ7's, V-19 and V-20, and applied to the grid V-15.

The reference voltage for the final regulating amplifiers is n from the RCA 5651 gas reference tube, V-18, which is operated m across the doubly regulated 250 volt output. The reference voltage is taken from the voltage divider R-35, P-5, and R-36 ough the two section R-C low pass filter to remove any rapid age fluctuations in the gas tube.

To provide maximum regulator stability V-14, V-15, V-19, V-20, , and V-17 are operated with d-c heaters fed from the regulated filament supply. Bishop and Harris point out[1.3] that stabilion of the heaters of the shunt amplifiers is necessary for a B+ valent signal of less than 50 microvolts.

The variable resistors P-4 and P-6 were used in setting up the uit and require no further adjustment. To adjust the output ages P-2, P-3 and P-5 were used. With the grid of V-12 conted to a +150 volt supply the output of the first regulating ion was adjusted to 350 volts by varying P-2. By varying P-5 P-3 together the -100 volt and +150 volt outputs were adjust-

With this regulator all currents are returned through the -100 volt lead and there is no current return through ground. For various of the d-c amplifier output circuits and for the plate supply of the heater regulator amplifiers a grounded high voltage supply is required. A 600 volt center-tapped transformer, T-4, a 5Y3 rectifier, and a single section capacitor input filter provide a +255 volt unregulated supply at 45 milliamperes with the negative side grounded.

PERFORMANCE

Rejection Ratio

Since it directly determines the useful sensitivity of the amplifier for measurements in the presence of interfering pick-up, the rejection ratio of the amplifier is of considerable importance. The term is used here in a different sense from the normally used definition of ratio of differential gain to in-phase gain. A voltmeter was connected between one oscilloscope output terminal and ground, and the output was measured for a balanced input signal and for a signal impressed upon both inputs tied together. The rejection ratio is then obtained by

$$\frac{e_{in}}{e_{diff}} \times \frac{E_{diff}}{E_{in}}$$

where e_{in} is the in-phase input signal, e_{diff} is the differential input signal, E_{in} is the output signal for the in-phase input, and E_{diff} is the output signal for the differential input. It is thus seen that the rejection ratio is, in effect, the factor by which the interfering in-phase signal can extend the differential signal for equal output indications.

To obtain the curves of Fig. 4 one of the amplifier channels (left amplifier) was adjusted for maximum ratio at the 50 microvolt sensitivity setting and for a frequency of 100 cycles. While the data was taken the only amplifier changes were an adjustment of the sensitivity and fine gain controls to obtain stated sensitivity and re-adjustment of the balance controls as required. For the 50 and 250 microvolt and 1 millivolt curves an in-phase signal of 1 volt was used; for the 10 millivolt curve the input signal was 2 volts and for the 100 millivolt curve it was 5 volts. The higher values were needed to obtain measureable output signals.

Since the μ-balance is adjusted to minimize the amplifier output the rejection ratio is a function of all the stages and thus of all circuit parameters. As expected, therefore, the rejection ratio changes with the sensitivity setting. The higher input signal lowers the rejection ratio as shown on Fig. 4 in two ways: 1) it changes the d-c potentials of the μ -balanced stage and thus unbalances it and 2) it contributes a larger in-phase signal output which is measured in addition to the amplifier generated differential output. The rejection ratio for the 100 millivolt sensitivity can thus be expected to be considerably greater for the usual range of in-phase interference signals.

Variation of the rejection ratio with frequency is to be ex-

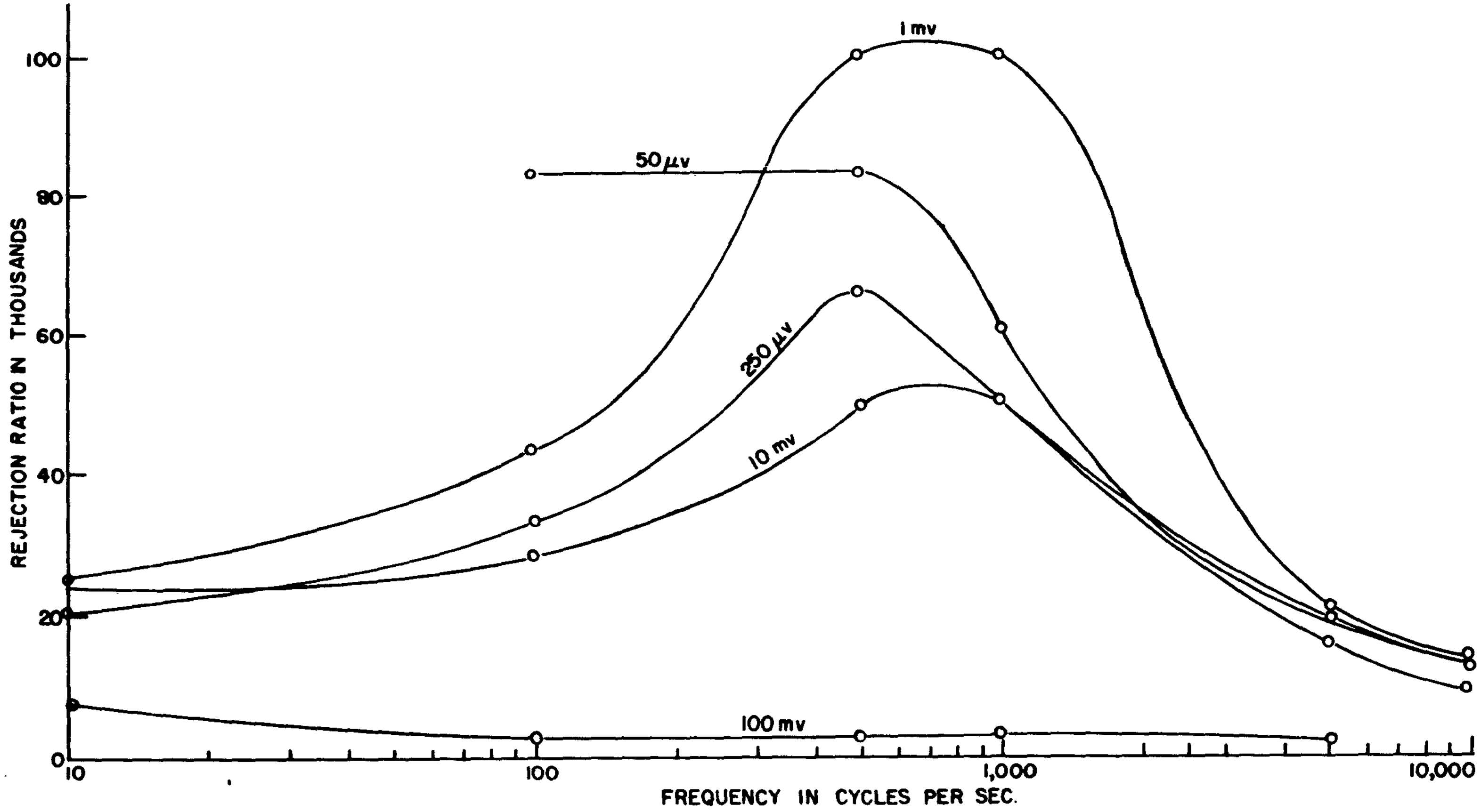


FIG.4 EFFECT OF FREQUENCY ON REJECTION RATIO

pected since there is no adjustment for unequal capacitances in the μ -balanced stage.

Frequency Response and Phase Shift

The desired goal was a response essentially flat to 10 kilocycles. The curves of Figs. 5 and 6 give the response obtained for each of the two amplifier channels. The differences in the responses for the two amplifiers result from slightly different adjustments in the coupling circuit and feedback circuit capacitors. As evidenced by the curves, the response is down 1 db at approximately 20 kilocycles and the 3 db attenuation occurs beyond 40 kilocycles for each amplifier at any sensitivity setting.

If special work should require a better high frequency response, the amplifiers could be re-adjusted to increase the high frequency response at any one amplifier sensitivity setting.

At the same time that the frequency response data was taken, information for the phase shift curves of Figs. 7 and 8 was obtained. At 10 kilocycles the phase shift is seen to range from 10 to 23 degrees depending upon the sensitivity. Although this is a large amount of phase shift, the phase shift over the region from d-c to 30 kilocycles is very closely proportional to frequency and thus introduces negligible waveform distortion.

The rise time of the amplifier was not accurately measured but was determined by the use of a square wave to be less than 20 microseconds.

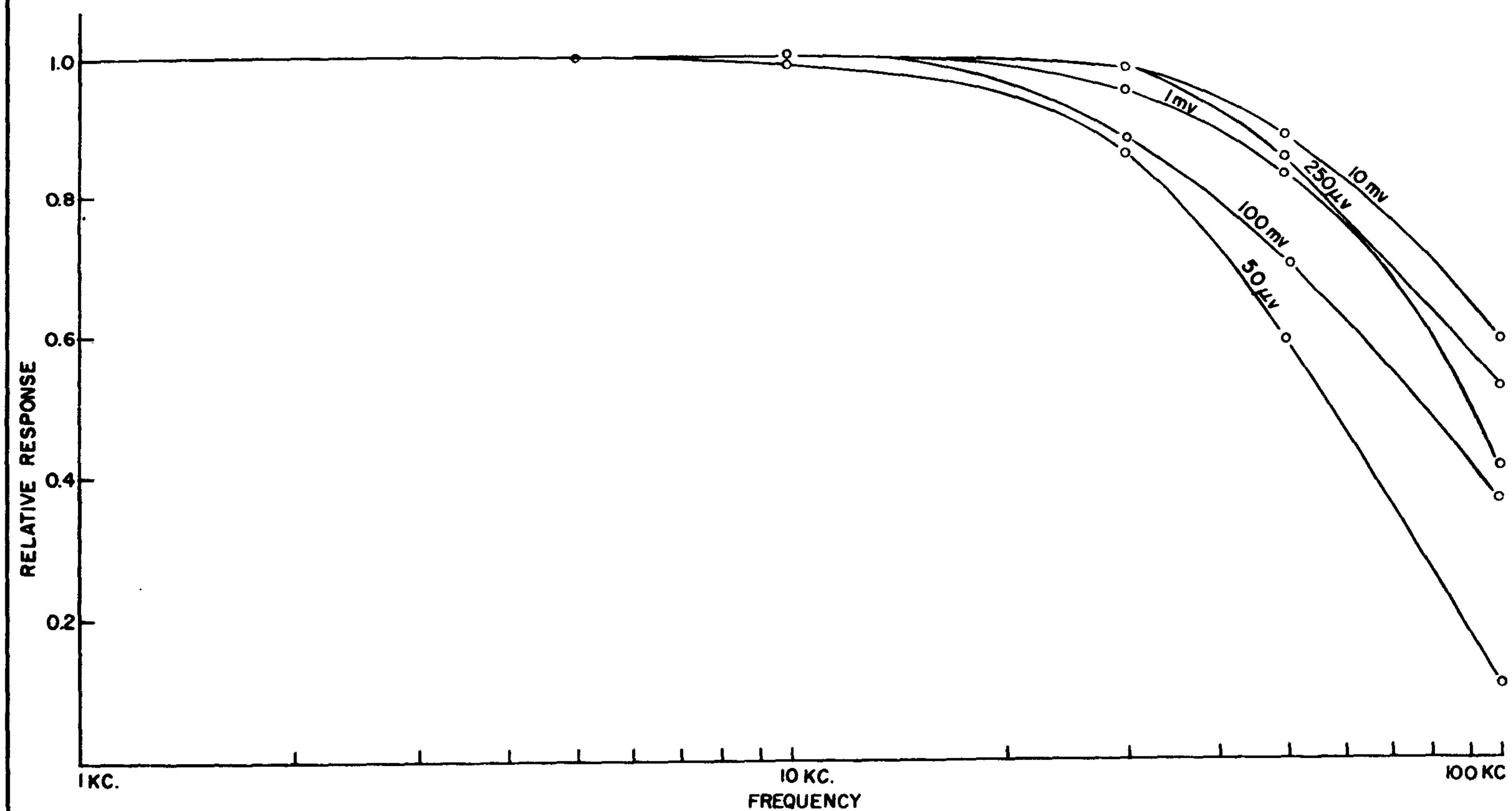


FIG. 5 FREQUENCY RESPONSE– RIGHT AMPLIFIER

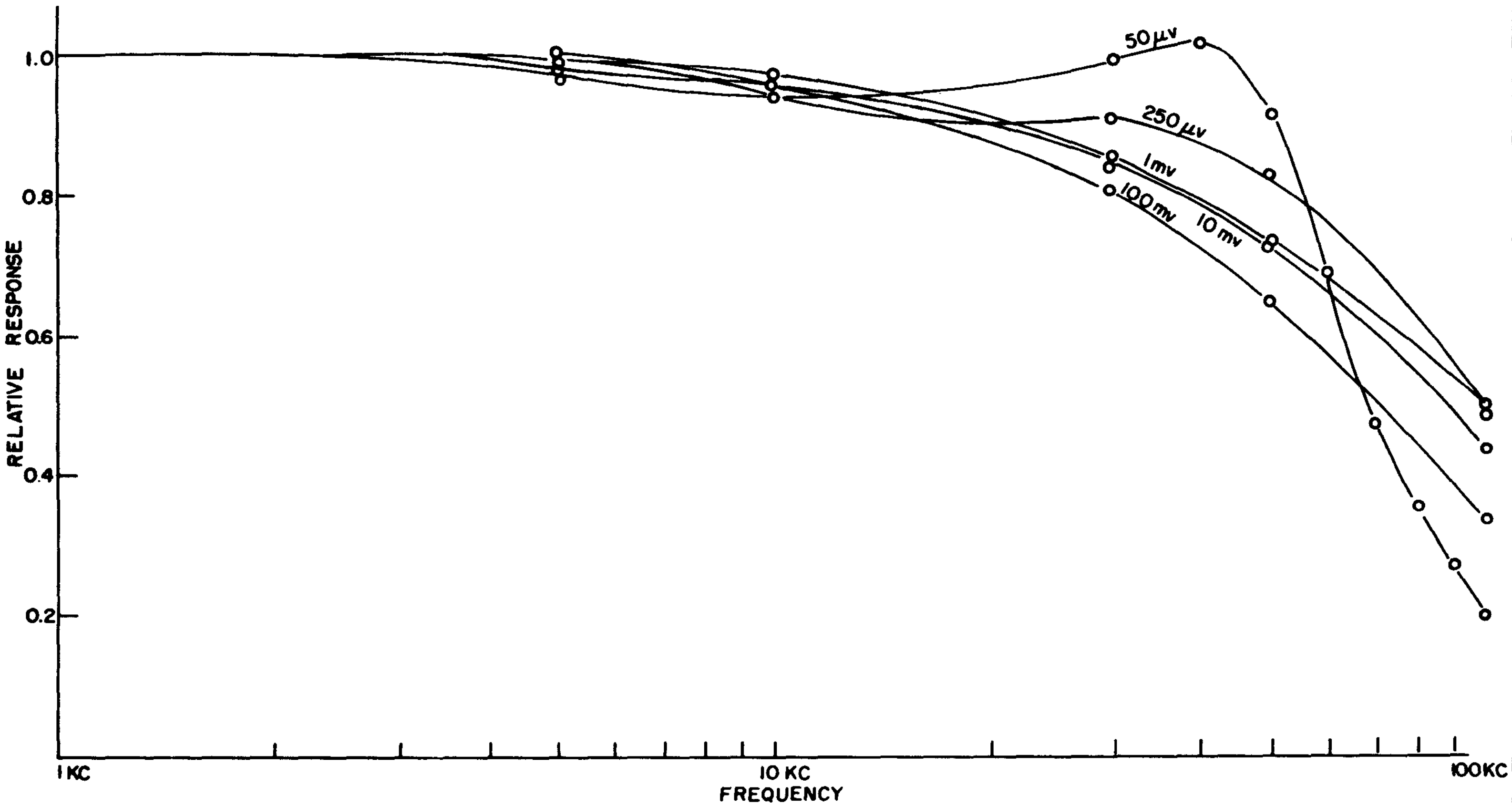


FIG.6 FREQUENCY RESPONSE – LEFT AMPLIFIER

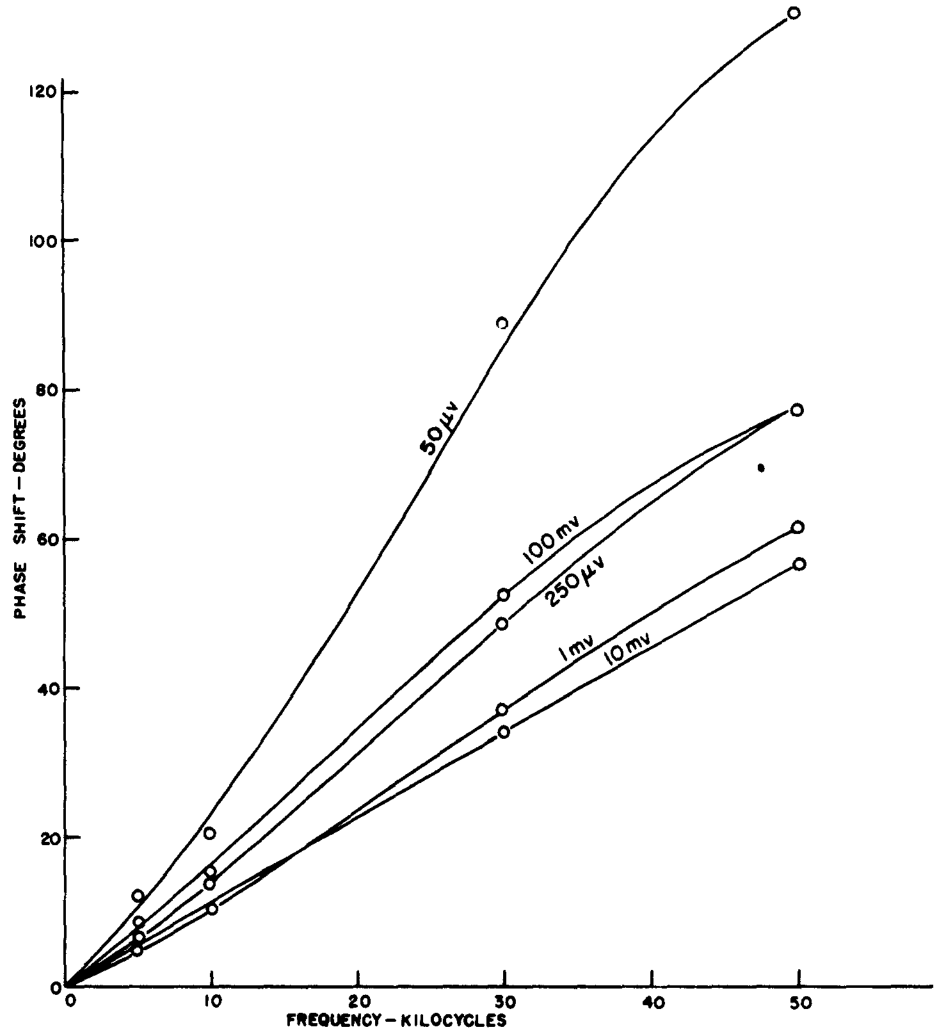


FIG.7 PHASE SHIFT-RIGHT AMPLIFIER

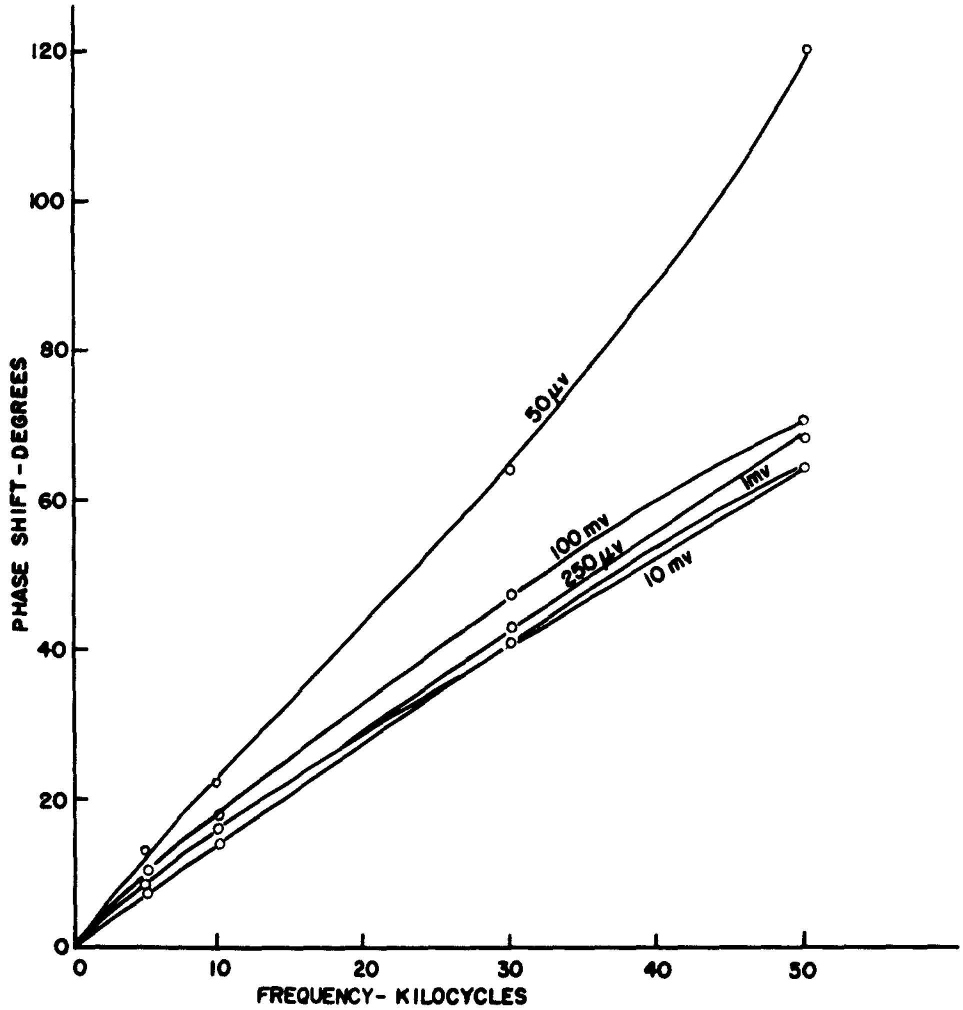


FIG. 8 PHASE SHIFT— LEFT AMPLIFIER

Input Impedance and Noise

The input resistance is limited by grid to ground resistors to 20 megohms for a balanced signal and 10 megohms for an unbalanced signal with one grid grounded. The measured input capacitance with approximately 2-1/2 foot shielded leads was measured to be less than 1.5 micromicrofarads between any input lead and ground.

The grid current measured flowing through the 10 megohm input grid resistors was between 4 and 5.3×10^{-10} ampere.

At the maximum bandwidth the rms noise output referred to the input grids is between 4.8 and 5.5 microvolts. This noise was measured with the input grids grounded and is free of any 60 cycle pick-up.

Stability

Amplifier stability records were taken with an electronic regulator between the power line and the amplifier. On the one millivolt sensitivity range a drift of less than 1.5 millivolts in 2.5 hours was recorded. At a sensitivity of 250 microvolts a peak-to-peak drift of 250 microvolts was measured over a one hour period. With a sensitivity of 50 microvolts a peak-to-peak drift of 100 microvolts was measured during a one-half hour period; the short time drift is, of course, much higher than the average drift rate.

Representative stability records taken with an Esterline Angus recording milliampere (1 milliampere full scale) are given in Figs. 9, 10, and 11.

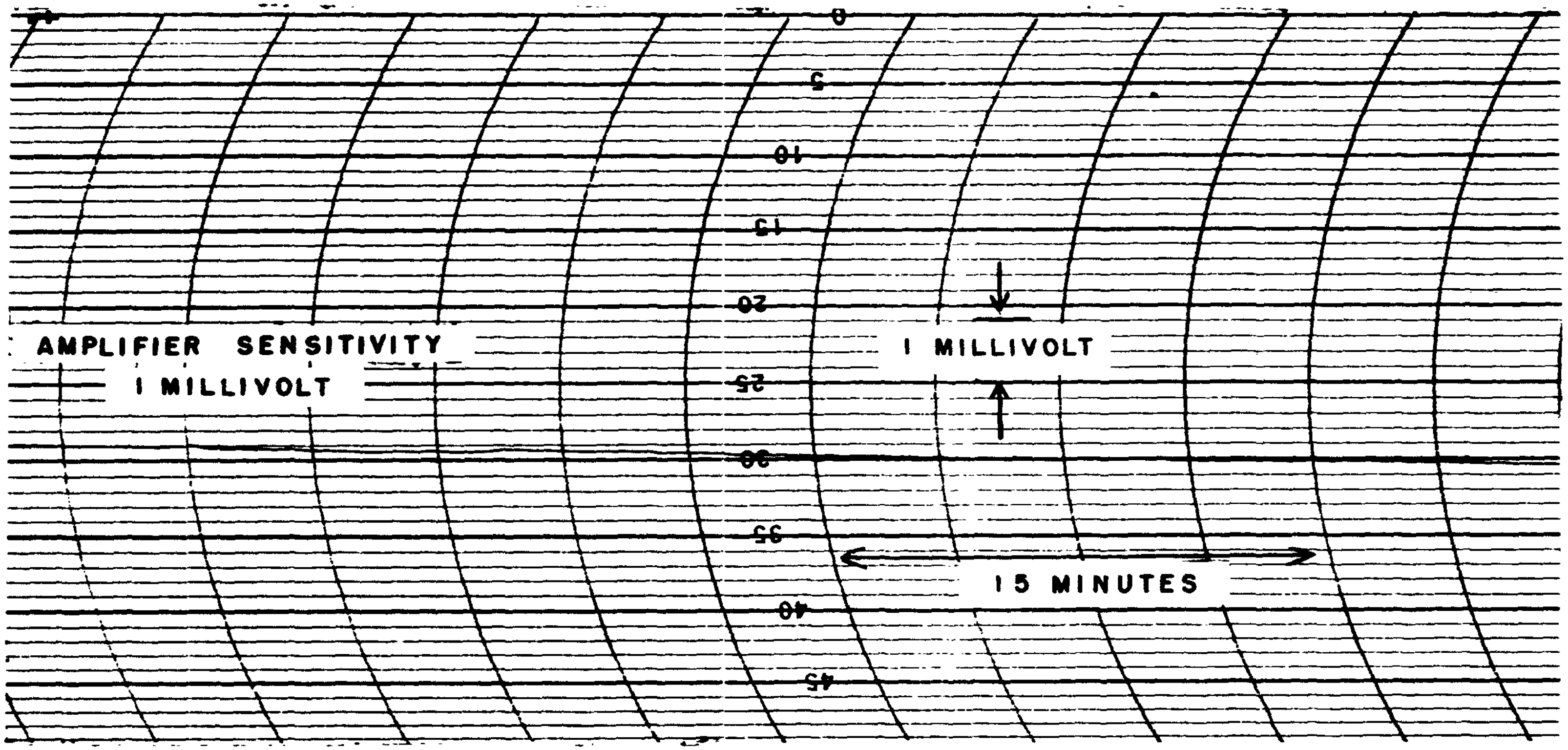


FIG.9 AMPLIFIER STABILITY RECORD

AMPLIFIER SENSITIVITY . 250 MICROVOLTS

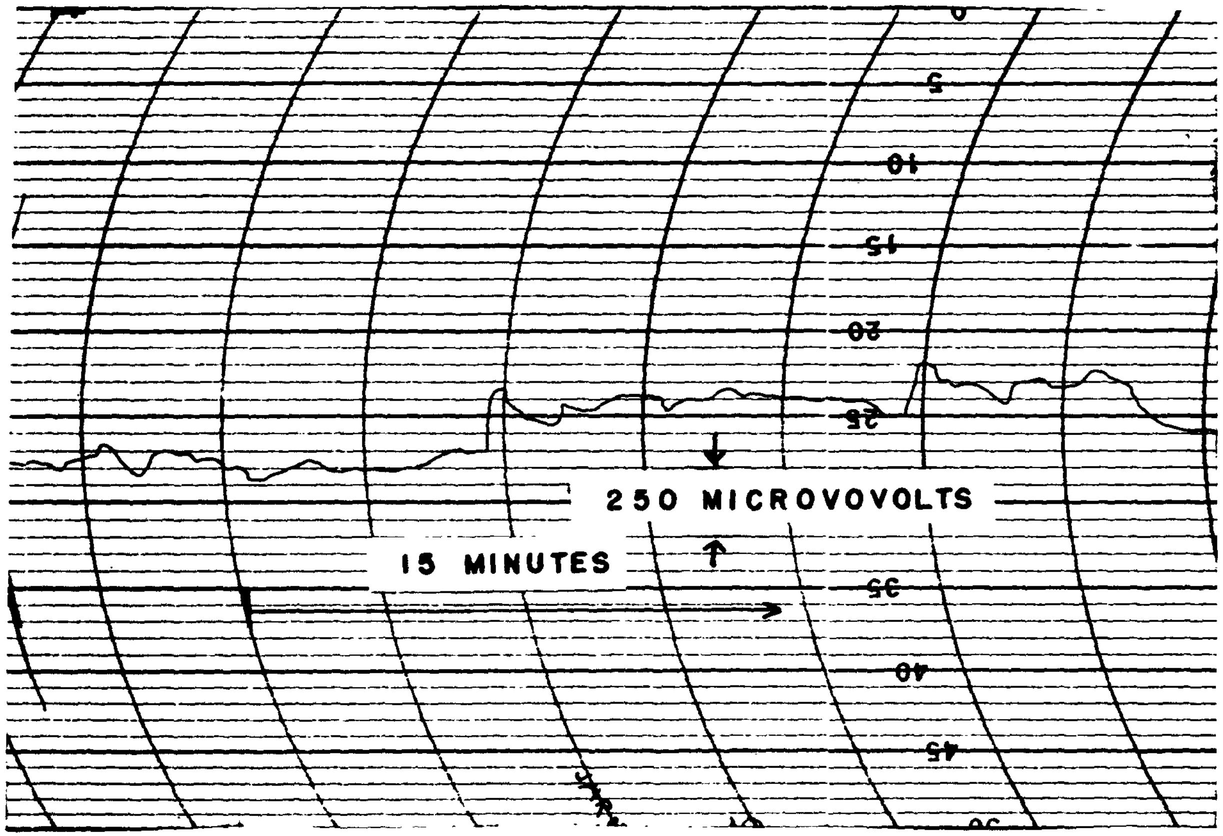


FIG. 10 AMPLIFIER STABILITY RECORD

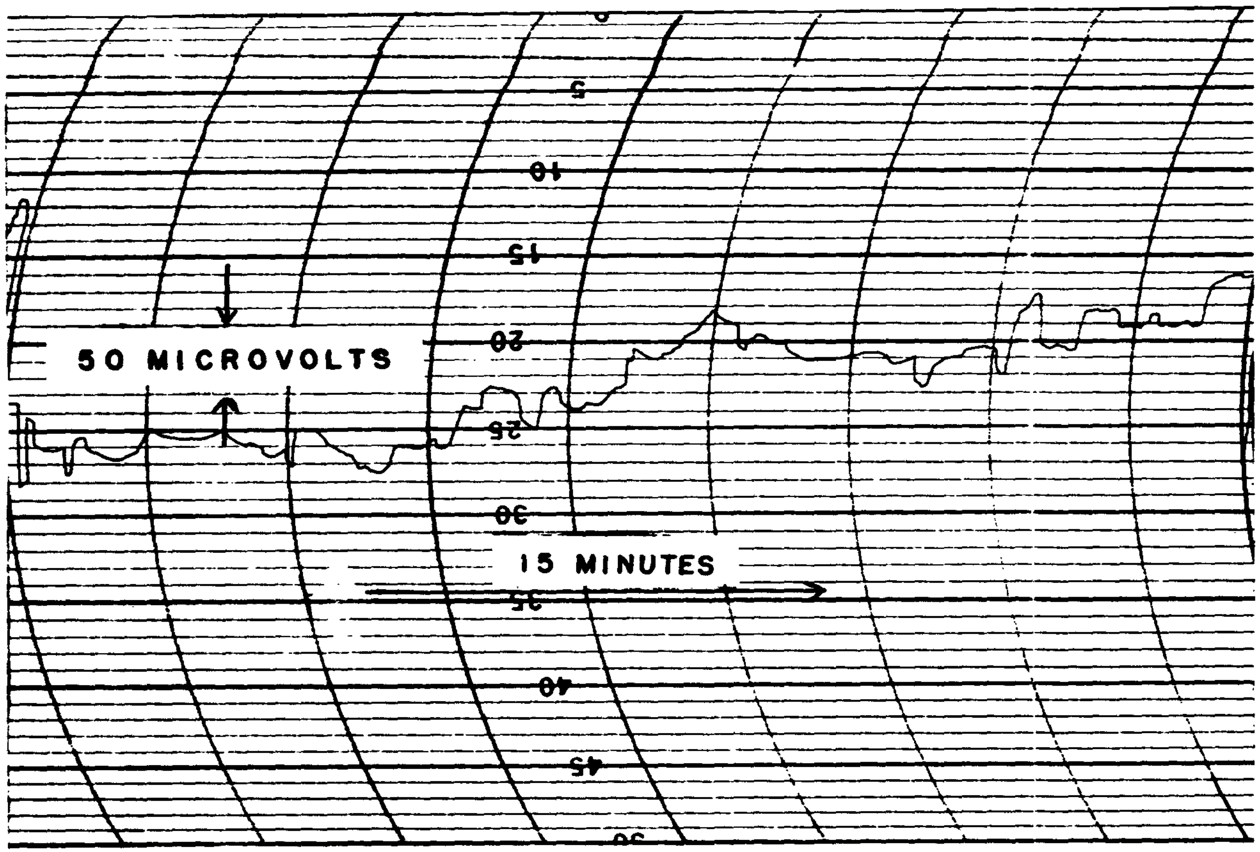


FIG. 11 - AMPLIFIER STABILITY RECORD

The effects of sudden changes in the output voltage of the power line regulator are shown in Figures 12 and 13. Fig. 12 was run with an amplifier sensitivity of 1 millivolt and the record indicates that a 5 volt change in the a-c supply to the amplifier represents an equivalent input signal of about 0.2 millivolt or less. Fig. 13, taken with a sensitivity of 250 microvolts, indicates approximately the same 200 microvolt equivalent input shift for a change in a-c supply of 5 volts. Since the amplifier is intended to be supplied from an electronic regulator whose stability is 0.1% of nominal (110 volts) the drift or spurious output signals do not arise from insufficient power supply stabilization ratios.

The performance of the heater power supply is illustrated by Fig. 14. The stabilization of the output is approximately constant for supply voltages from 105 to 125 volts. The output falls off rapidly below 105 volts because the load regulating 6L6's reach cut-off bias at about this point and the regulating circuit becomes inoperative. In the region of regulation the output changes are reduced to 0.006 volt per volt change in a-c supply voltage.

Interaction Between Amplifier Channels

When one amplifier channel is set at minimum sensitivity and a 100 millivolt differential signal is applied to the input, there is no discernible output in the other channel set at a 50 microvolt sensitivity with the input grounded.

SENSITIVITY SETTING. 1 MILLIVOLT

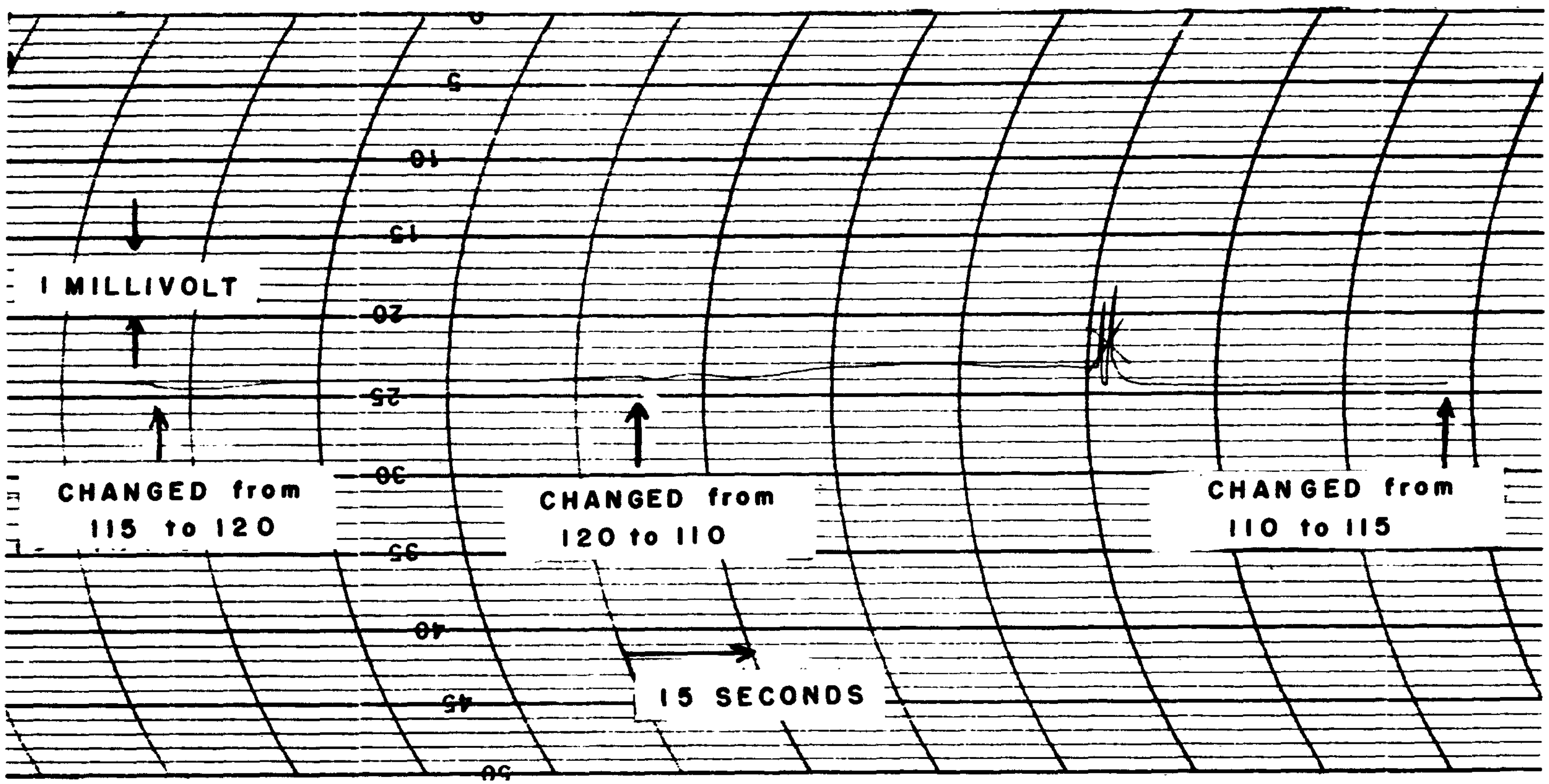


FIG.12. EFFECT OF RAPID SHIFT IN AMPLIFIER A.C. SUPPLY VOLTAGE

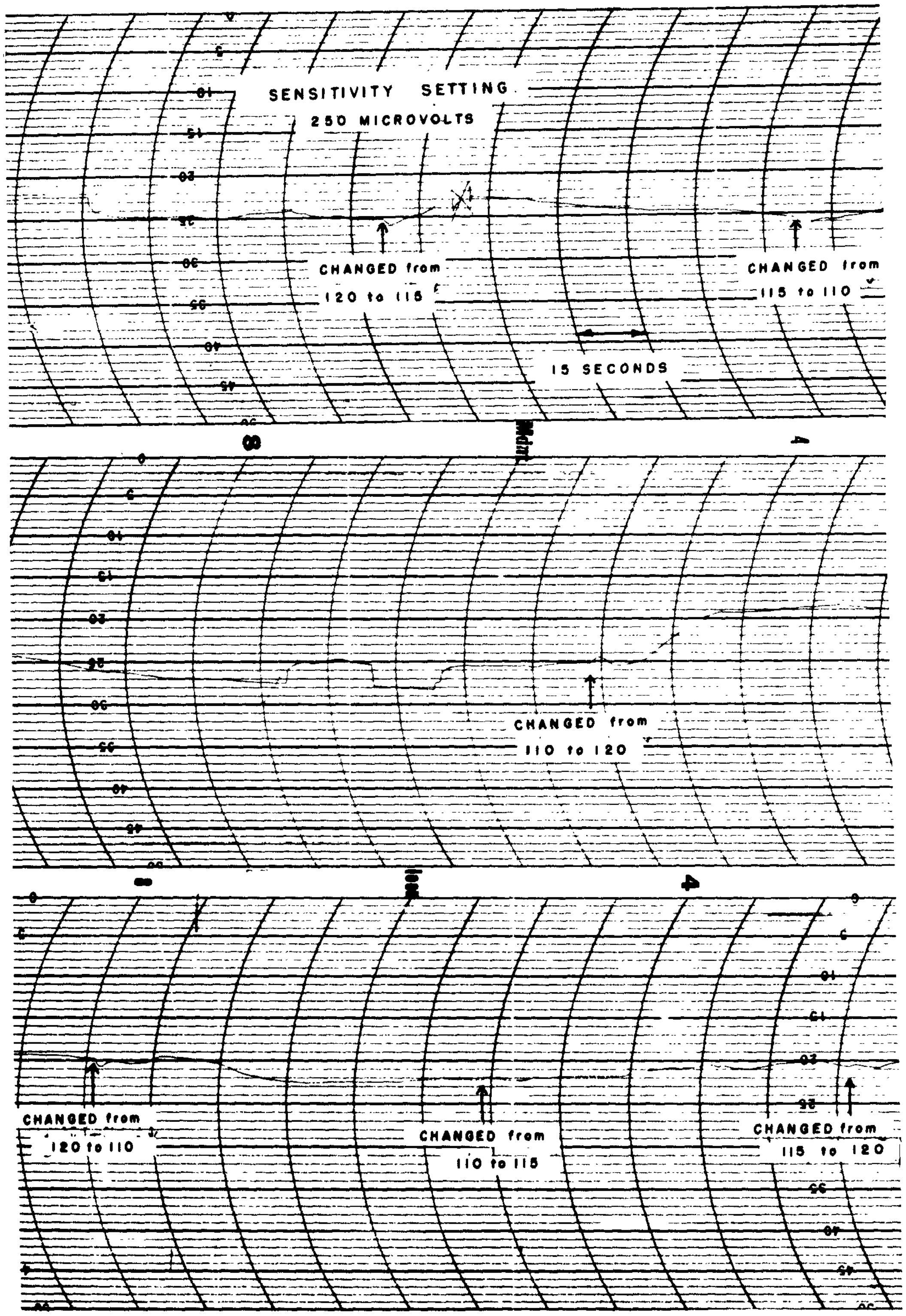


FIG. 13 EFFECT OF RAPID SHIFT IN AMPLIFIER A.C SUPPLY VOLTAGE

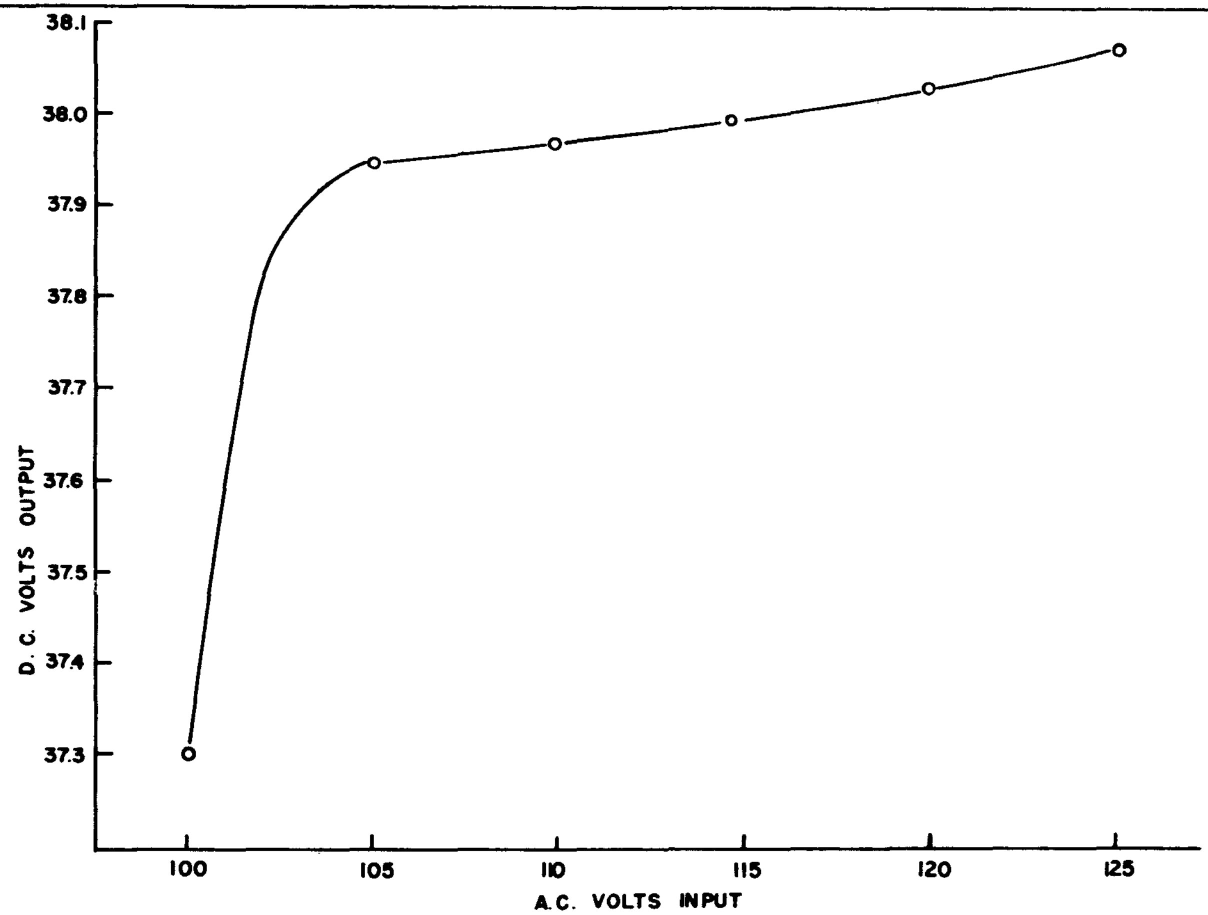


FIG. 14 STABILITY OF HEATER SUPPLY REGULATOR

EQUIPMENT DESCRIPTION

The amplifier with the associated power supplies consists of four chassis mounted on standard 19-1/2 inch rack mounting panels assembled in a 38 inch high cabinet as pictured in Fig. 15. A rear view of the equipment showing the interconnecting cables is pictured in Fig. 16.

The lowest unit, the Power Supply Chassis, contains all the large transformers and chokes, the heater supply circuits, and the heater regulating 6L6's. These large a-c current devices are located at the bottom of the cabinet to isolate them from the amplifier and thus minimize possibilities of pick-up. The only control on this chassis is the power switch on the panel; in addition, the pilot light and the power fuse are mounted on the panel.

The Regulator Chassis is mounted above the Power Supply Chassis and contains no controls nor indicators on the panel. This chassis contains all the remaining power supply circuits including the entire unregulated + 255 volt power supply.

The main Amplifier Chassis, mounted above the Regulator Chassis, contains two identical amplifying channels which are constructed as mirror images. As pictured on Fig. 15 each amplifier contains four main operating controls on the front panel. At the top of the panel are the two gain controls; the five position SENSITIVITY switch and the FINE GAIN control. Directly below the fine gain control is the BALANCE adjustment, and below the sensitivity switch is the OUTPUT ADJUST potentiometer used for balancing the final amplifier stages.

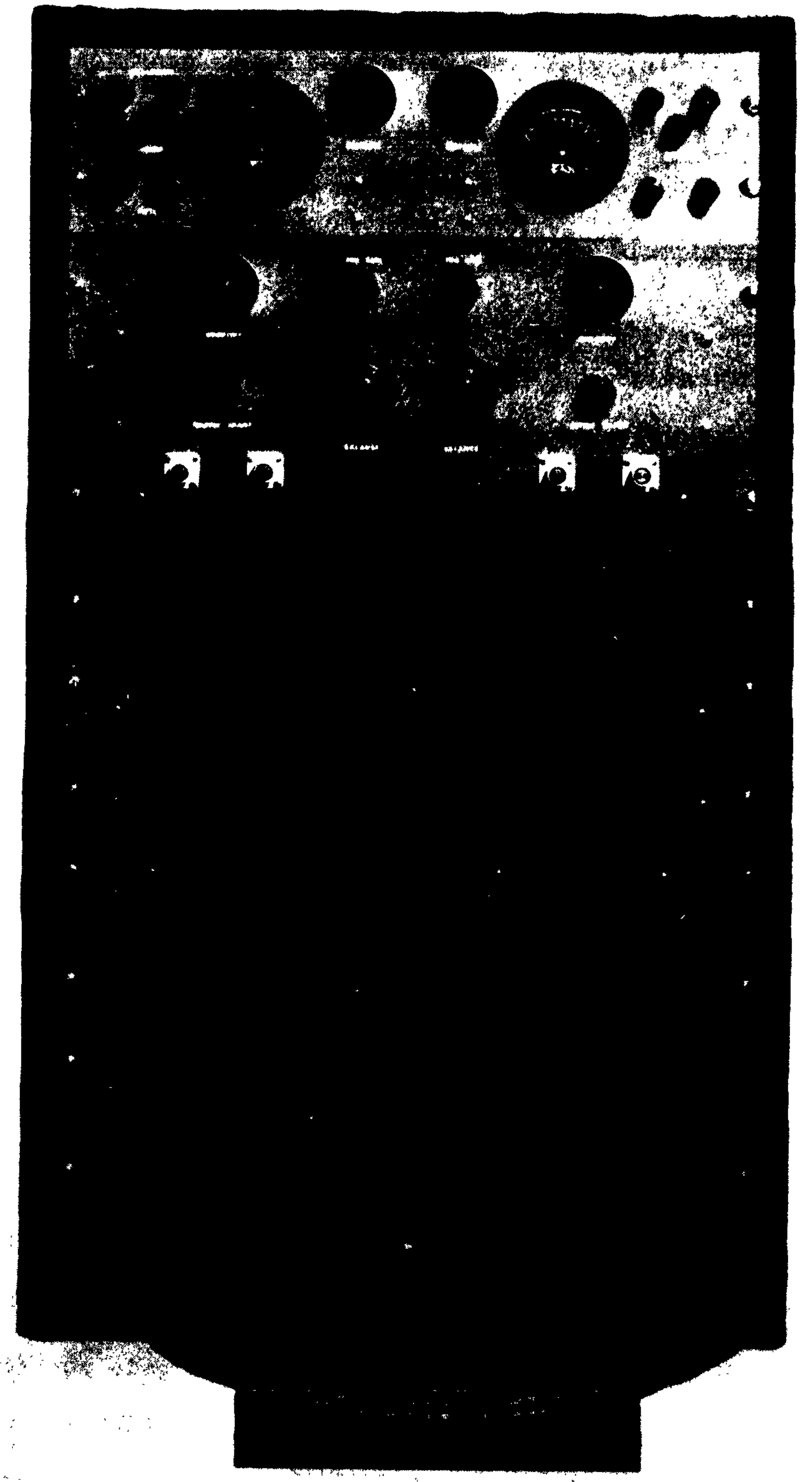


FIG. 15 AMPLIFIER-FRONT VIEW

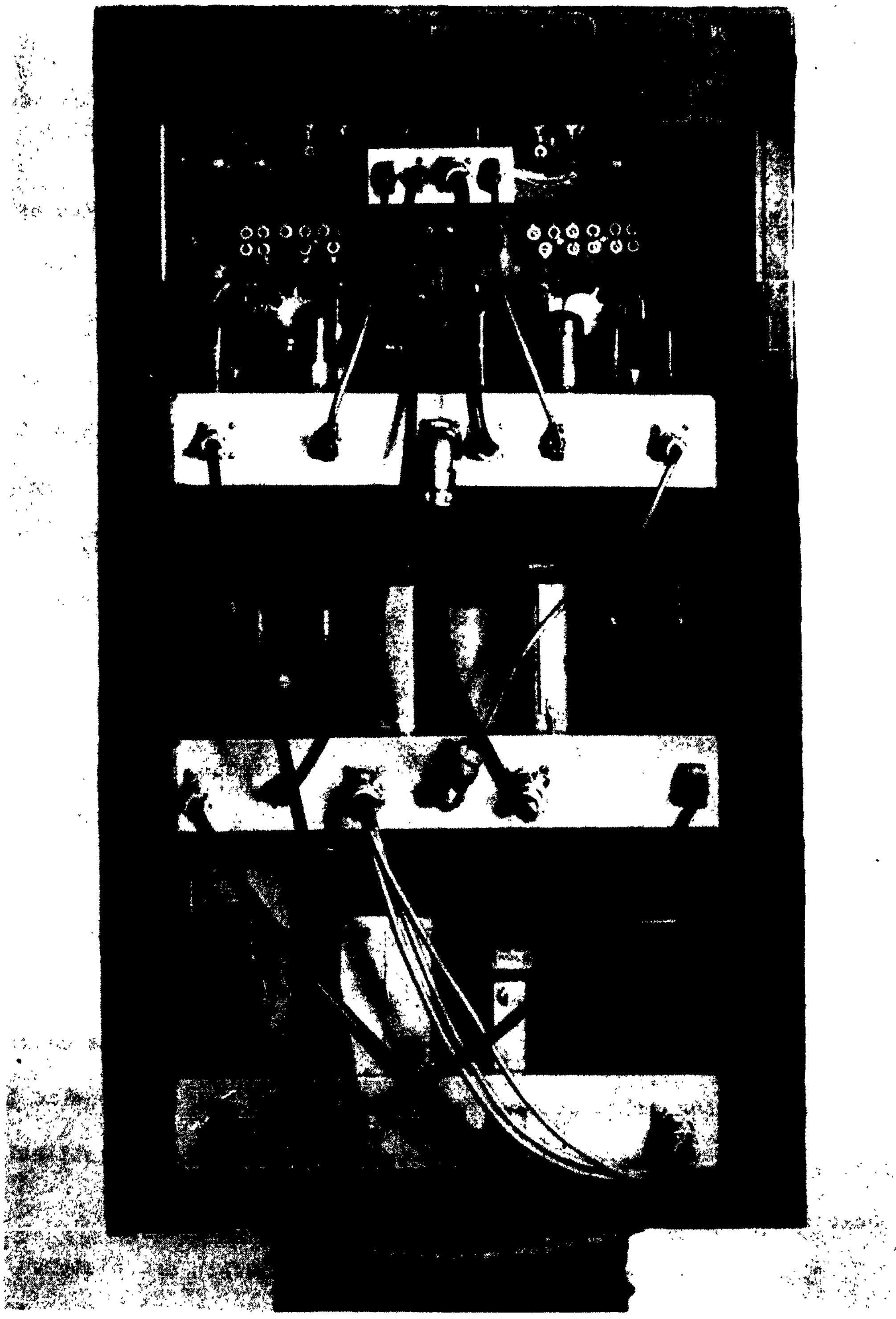


FIG. 16 AMPLIFIER-REAR VIEW

At the bottom of the panel are four input connectors - two for each amplifier. Each connector is numbered to correspond with the numbered input cable (not shown) for which the amplifier has been balanced.

The final panel holds the Output Chassis which is seen to be a small chassis to permit access to the Amplifier Chassis from the lid in the top of the cabinet. This chassis contains two identical output circuits constructed as mirror images. Each half of the panel contains a balance indicating meter, a five position RESPONSE switch which selects the upper frequency limit, and five output terminals. The lower METER pair of terminals is at approximately +100 volts while the two OSCILLOSCOPE terminals are at ground potential. An additional GROUND terminal is provided which is connected to the grounded circuit.

Because of the large amount of heat to be dissipated the cabinet back was removed and the top lid was replaced with a grillwork lid.

Figs. 17 and 18 give a top and bottom view of the Power Supply Chassis with the various components labeled for identification. Capacitors C-4 and C-5 should preferably be mounted above the chassis but were mounted below because of space limitation. To provide for cooling of R-1, which dissipates approximately 50 watts, ventilating holes were drilled along the side of the chassis.

Top and bottom views of the Regulator Chassis are pictured in Figs. 19 and 20. The six regulator amplifiers are mounted on a

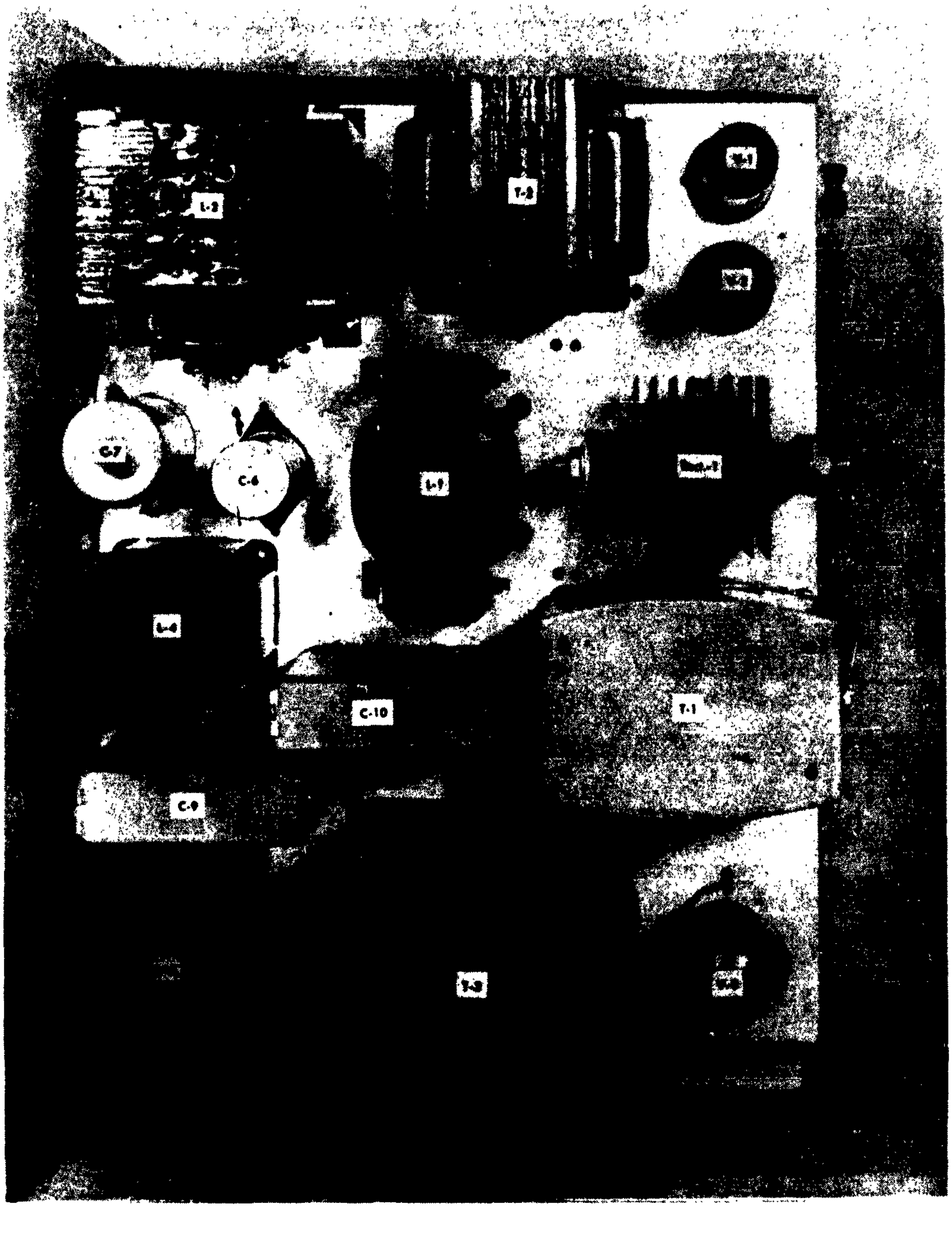


FIG. 17 POWER SUPPLY CHASSIS-TOP VIEW

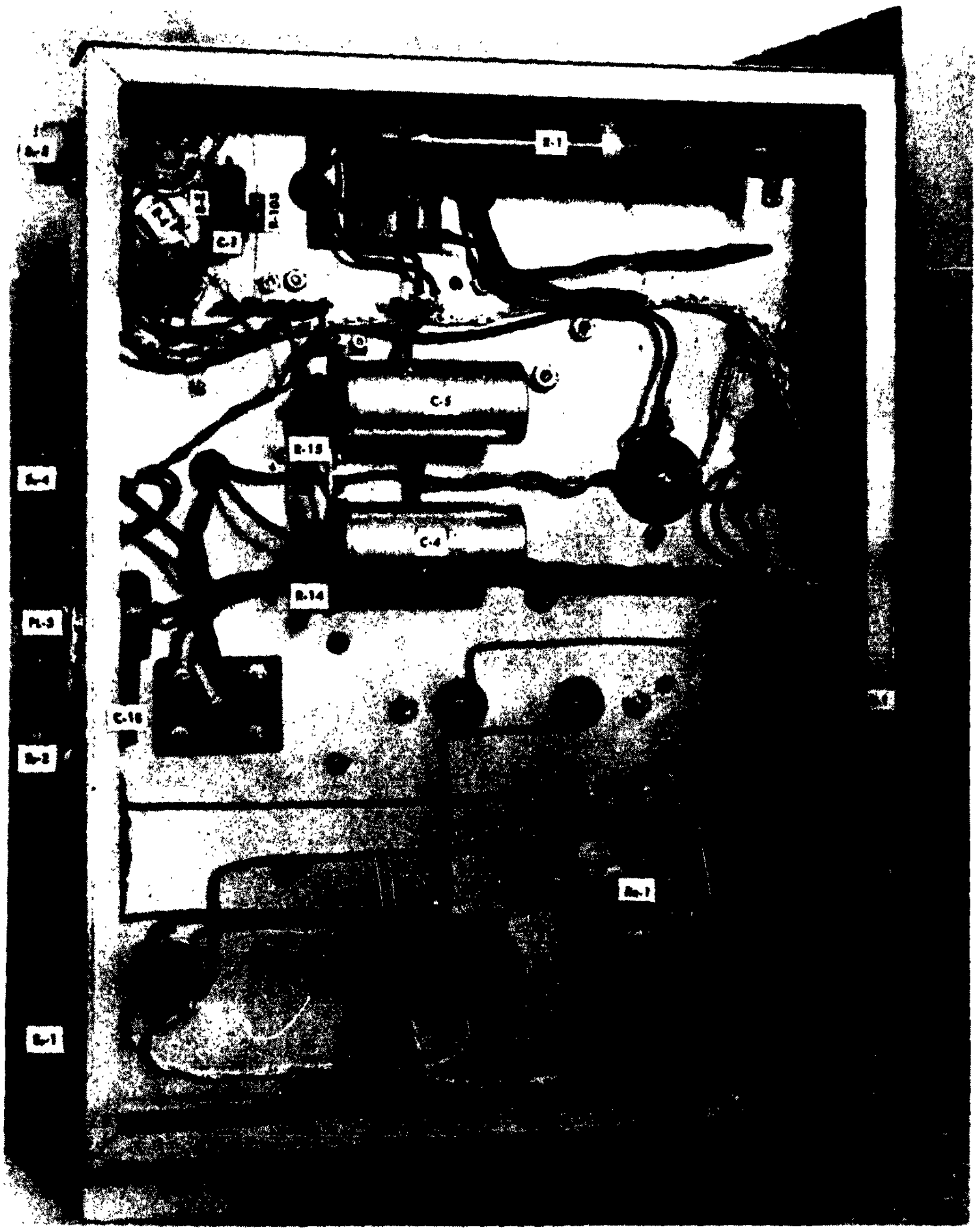


FIG. 18 POWER SUPPLY CHASSIS- BOTTOM VIEW

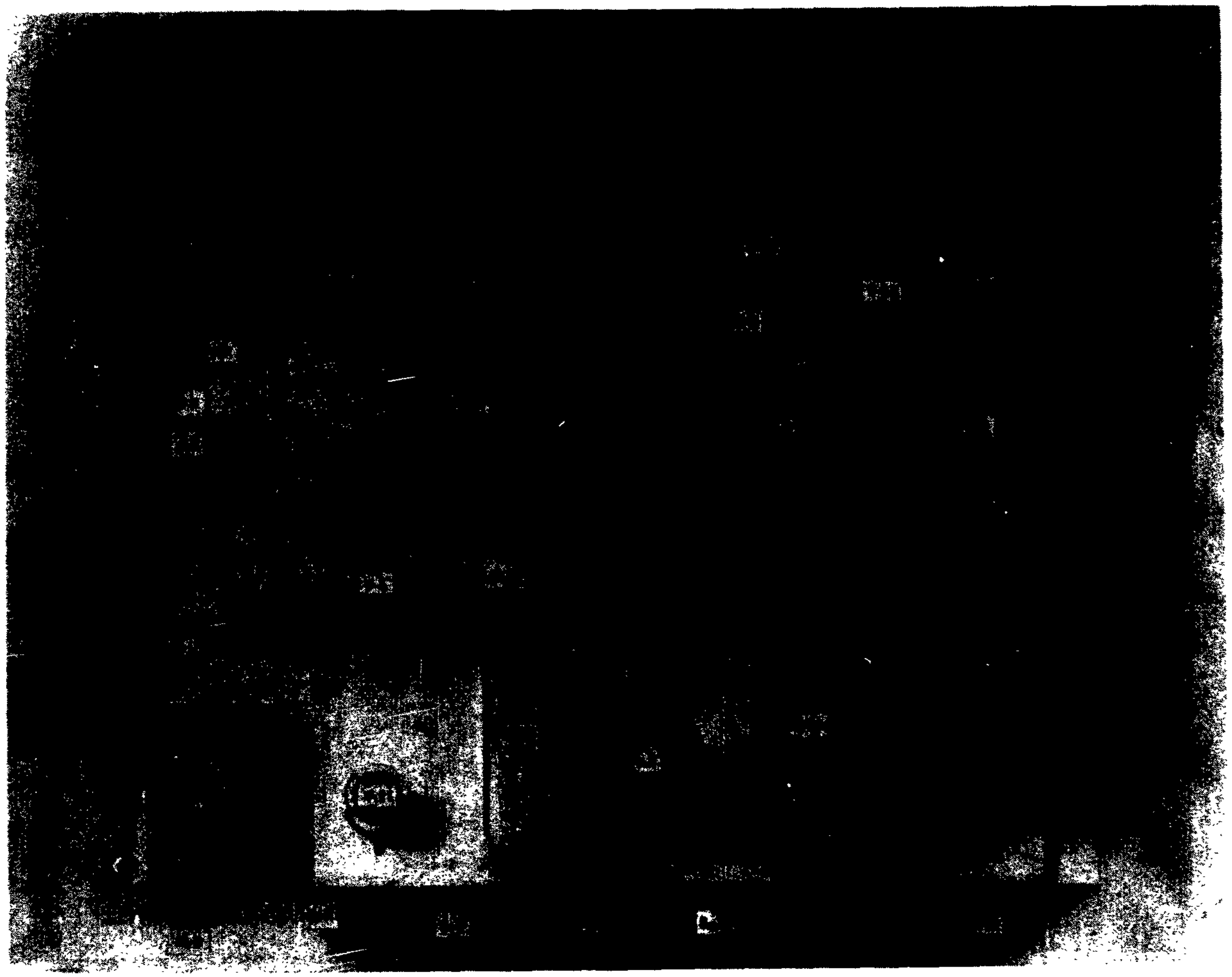


FIG. 19 REGULATOR CHASSIS-TOP VIEW

FIG. 20 REGULATOR CHASSIS-BOTTOM VIEW

sub-chassis shock mounted with Barry vibration mounts. Connections between the sub-chassis and the main chassis were made with extremely flexible wire to maintain the vibration isolation. Although all adjustments are made available from the top of the chassis, it is not necessary to make any of these adjustments in normal operation of the amplifier.

The Amplifier Chassis is shown in Figs. 21 and 22. The sub-chassis, shock mounted with Barry vibration mounts, contains the input, first, and second stage tubes and the first stage plate supply cathode follower. The third stage, the feedback duo-triodes, and the pentode cathode load tubes use rubber grommet mounted tube sockets to provide vibration reduction.

The cluttered appearance of the under chassis results from point to point wiring, the use of extremely flexible and coiled wires to vibration mounted circuits, and from the fact that this chassis served as both breadboard and final model.

Except for the feedback circuit capacitors and the coupling circuit compensating capacitors all adjustments are available from the front panel or from the top of the chassis. Controls on the top of the chassis are not needed in normal operation of the amplifier - only when tube aging shifts unbalance outside the range of the panel controls. Although not discernible from the photographs the input balancing capacitor, C-36, can be adjusted through an access hold in the top of the sub-chassis.

In operation the feedback and gain varying resistor groups and

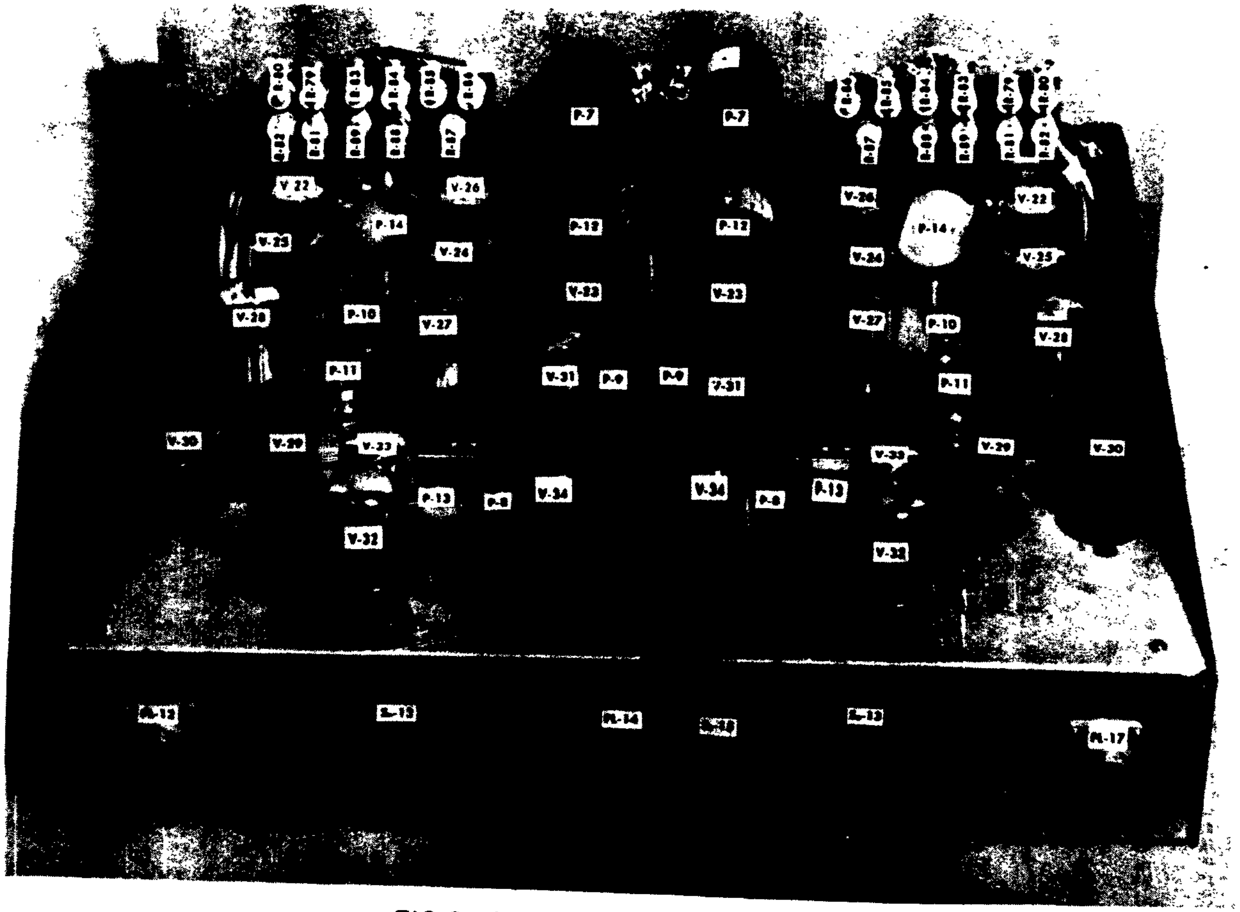


FIG. 21 AMPLIFIER CHASSIS-TOP VIEW

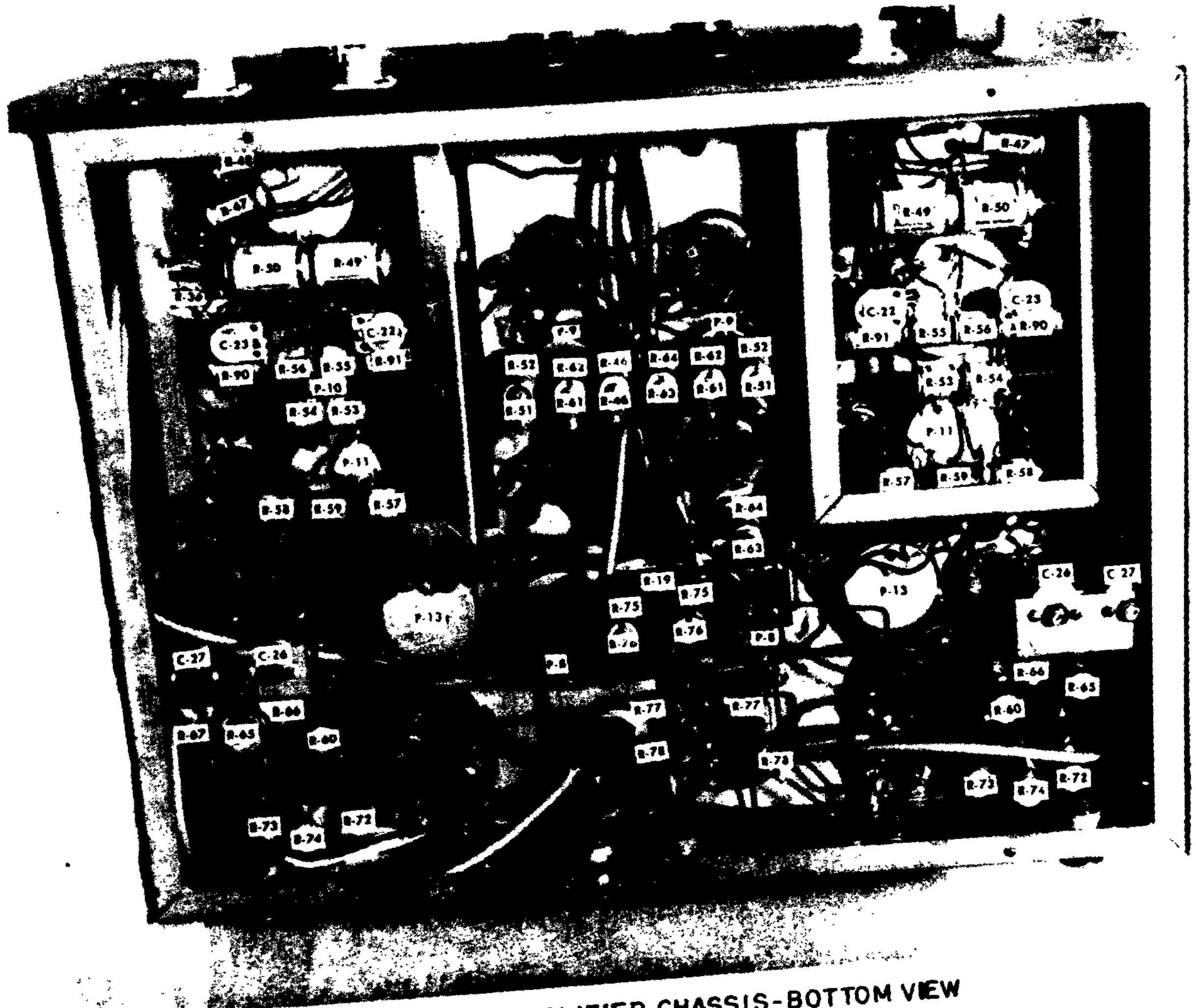


FIG. 22 AMPLIFIER CHASSIS-BOTTOM VIEW

the fine gain controls, P-7, are covered by a single galvanized iron shielding cover.

The Output Chassis shown in Figs. 23 and 24 permits easy access to the controls on the Amplifier Chassis mounted directly below it.

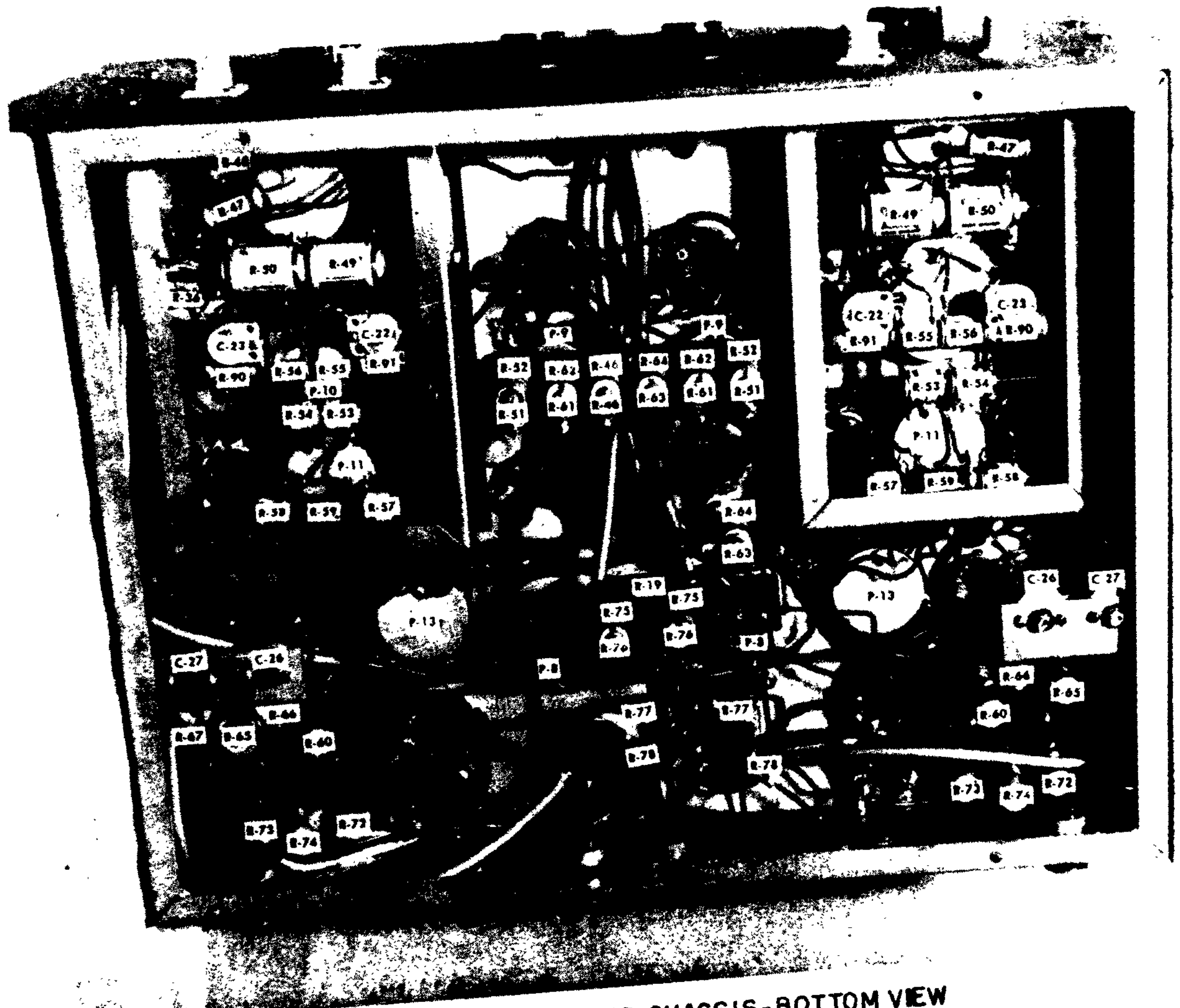


FIG. 22 AMPLIFIER CHASSIS-BOTTOM VIEW

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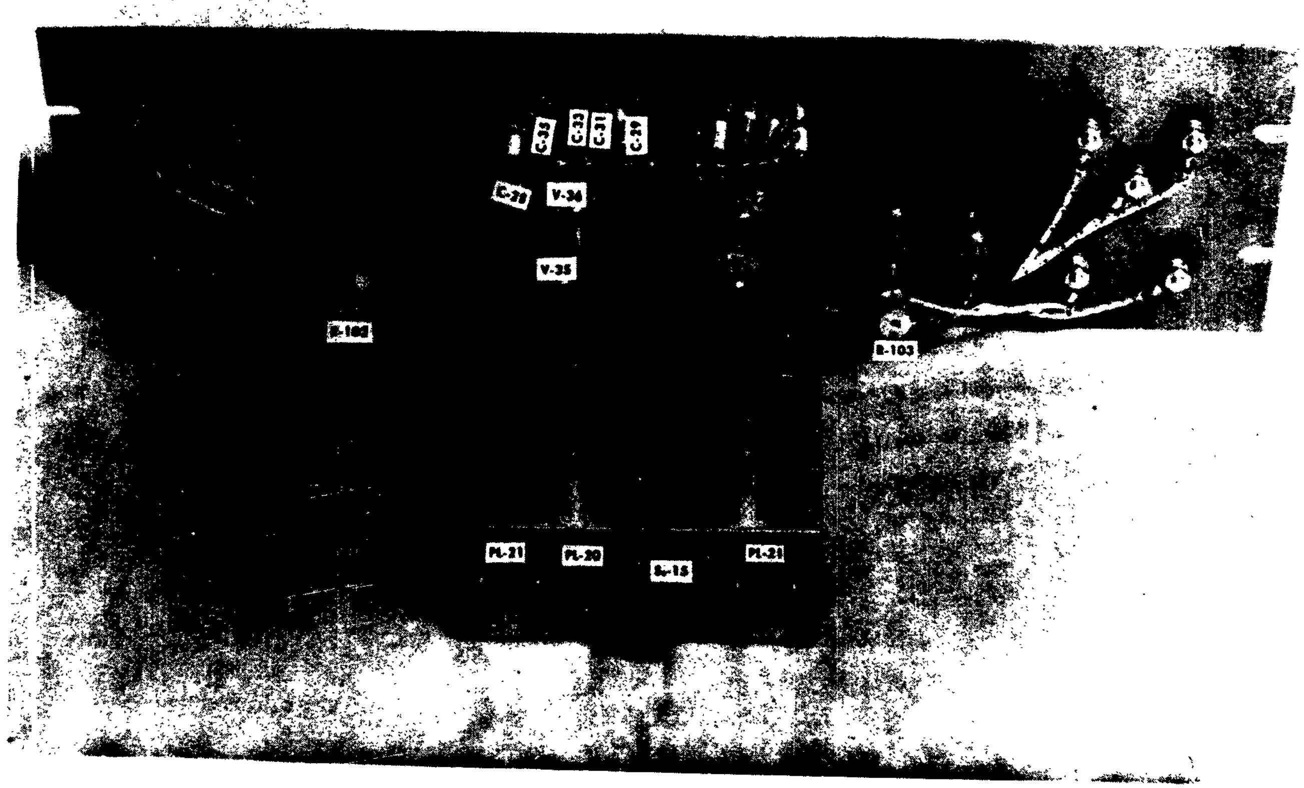

FIG.23 OUTPUT CHASSIS - TOP VIEW

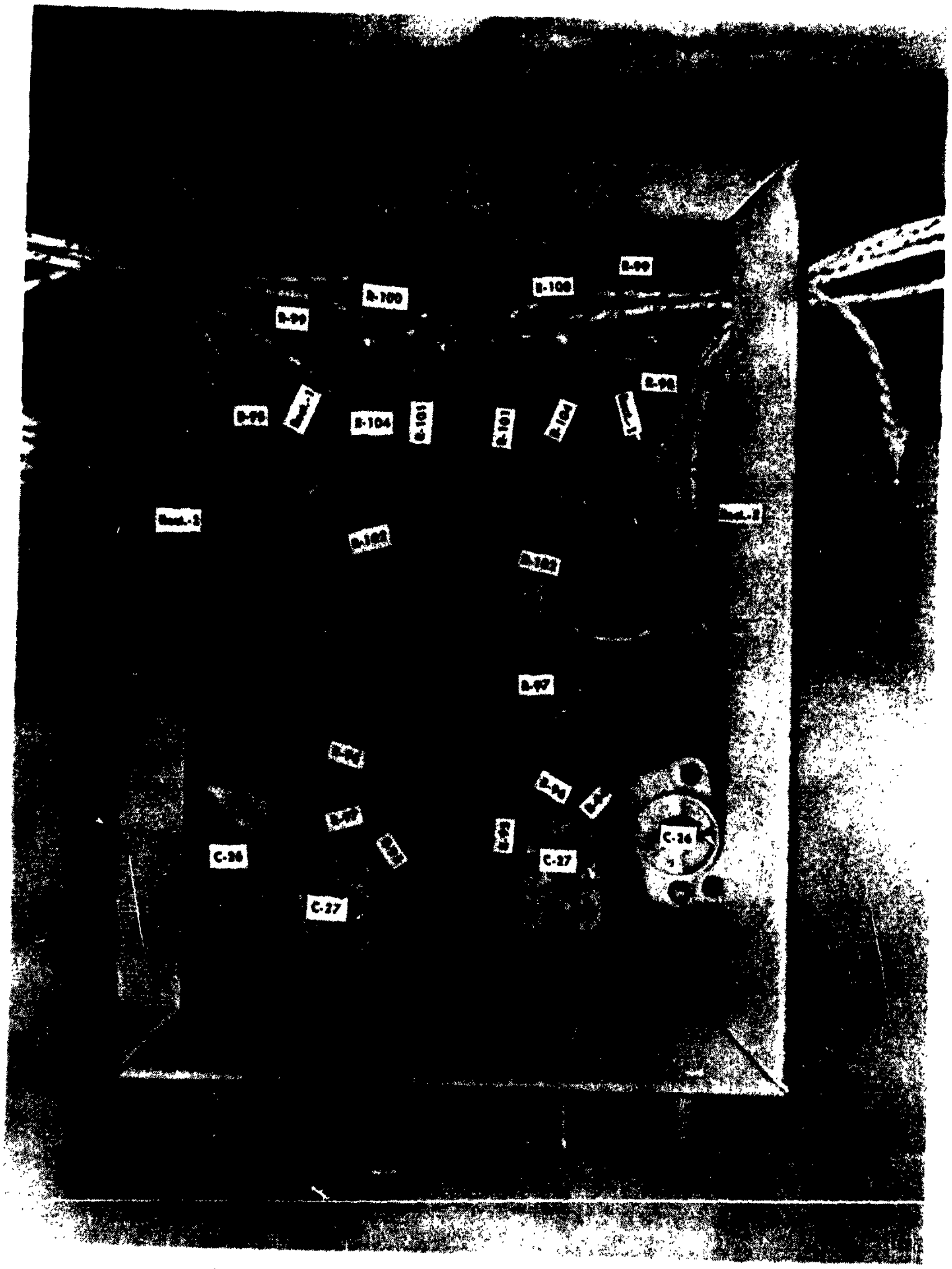

FIG. 24 OUTPUT CHASSIS - BOTTOM VIEW

APPENDIX I

Operating Instructions

1. Normal Operating Procedure

1.1 Preliminary

a) Set both SENSITIVITY switches to position 1.

b) Turn power switch ON.

c) Allow amplifier to warm up for at least 30 minutes before adjustment or use.

d) If the input cables have been removed, check to insure that the cables have been re-connected to their correspondingly numbered input connectors.

1.2 With no signal on the input leads (e.g. input leads grounded) adjust the panel meter to mid-scale (0.5 milliamperes) by using the OUTPUT ADJUST potentiometer.

1.3 Set SENSITIVITY switch to position 2 (even if position 1 will ultimately be used for measurement).

1.4 Balance the amplifier by bringing the panel meter to mid-scale by using the BALANCE potentiometer.

1.5 Set SENSITIVITY switch to the desired sensitivity position.

1.6 Repeat step 1.4.

1.7 Connect measuring instrument to the appropriate output terminals using either METER or OSCILLOSCOPE and GROUND terminals.

Note: If the instrument used for measuring the output voltage requires that one of its terminals be grounded, the GROUND terminal and one of the OSCILLOSCOPE TERMINALS must be used. Either OSCILLOSCOPE terminal may be used; their outputs are equal and of opposite phase. The average potential of the two OSCILLOSCOPE terminals is at ground potential. When using the METER terminals it should be remembered that they are at +100 volts above ground.

1.8 By using either OUTPUT ADJUST control (if SENSITIVITY switch is on position 1) or BALANCE control (if SENSITIVITY switch is on position 2, 3, 4, or 5) bring measuring instrument to zero.

1.9 Set the RESPONSE switch to the desired position.

Note: This control is effective only upon the OSCILLOSCOPE terminals.

1.10 Impress a calibrating signal upon the input leads of a magnitude commensurate with the sensitivity setting of the SENSITIVITY switch. Either note the deflection or adjust the amplifier gain to give a convenient deflection on the indicating instrument by adjusting the FINE GAIN control. It should be noted, however, that for an input signal equal to the nominal sensitivity of the amplifier at the particular SENSITIVITY switch position being used, the amplifier is designed to give an output of 1 volt between <u>either</u> OSCILLOSCOPE terminal and GROUND. The corresponding output at the METER terminals is 4.8 volts.

Note: 1) This calibrates or adjusts the amplifier only at this particular SENSITIVITY switch position. If the SENSITIVITY switch is changed to another position the amplifier must be recalibrated or the gain must again be adjusted with the FINE GAIN control.

2) During the gain adjusting procedure the amplifier balance should be checked and adjusted, if necessary, by using the BALANCE control or, if on SENSITIVITY position 1, by using the OUTPUT ADJUST control. This is necessary because of amplifier drift and because the change in the FINE GAIN potentiometer has a slight effect upon the amplifier balance.

1.11 In step 1.8 it may be noted that the measuring instrument and panel meter will not simultaneously indicate a balanced condition (i.e. zero output voltage at the output terminals or mid-scale on the panel meter). This is not an abnormal condition since the panel meter is only an approximate indication of balance. The measuring instrument is the one which should be brought to a zero reading as indicating a balanced amplifier. It should be noted further that the OSCILLOSCOPE and METER terminals will not necessarily be balanced simultaneously.

2. <u>Precautions</u>

2.1 When using a sensitive meter on the METER terminals do not connect the meter to the terminals if the panel meter does not read approximately mid-scale.

Note: From mid-scale (0.5 ma) a deflection of ± 0.1 ma indicates a potential differential across the METER terminals of approximately 5 volts.

2.2 To minimize the possibility of damaging a meter connected across the METER terminals it is necessary to turn the SENSITIVITY switch to position 1 whenever making any connections to the input leads.

3. Adjustments

3.1 If the range of the OUTPUT ADJUST potentiometer is insufficient to balance the amplifier in step 1.1 it is necessary to readjust the coarse control located on the amplifier chassis.

a) Be sure the SENSITIVITY switch is on position 1.

b) Set the OUTPUT ADJUST potentiometer in the center of its range (arrow vertical).

c) Bring the panel meter to mid-scale by adjusting potentiometer P-13.

3.2 If the amplifier cannot be balanced in step 1.4 by using the BALANCE potentiometer, the additional coarse control on the chassis must be used.

a) Set the SENSITIVITY switch to position 1 and bring the panel meter to mid-scale by using OUTPUT ADJUST control as in step 1.2.

b) Turn SENSITIVITY switch to position 2.

c) Set BALANCE control to 5.00 (center of its range).

d) Bring meter to mid-scale by adjusting potentiometer P-11.

3.3 If using the OSCILLOSCOPE and GROUND output terminals, it may be necessary to bring the average potential of the OSCILLOSCOPE terminals to ground potential. This adjustment is indicated if the oscilloscope, previously adjusted and using the d-c deflection amplifier, has a large deflection when the panel meter is at mid-scale. A small deflection can be compensated with the vertical position adjustment on the oscilloscope. A potentiometer on the amplifier chassis is provided for making the amplifier output level adjustment.

a) Set SENSITIVITY switch to position 1.

b) Place OUTPUT ADJUST control at the center of its range (arrow vertical)

c) With the oscilloscope input temporarily shorted bring the trace to center of screen. Remove short.

d) Bring panel meter to mid-scale by adjusting P-13.

e) Bring trace to center of screen by adjusting potentiometer P-8. This will probably cause the panel meter to move off mid-scale.

f) Re-adjust panel meter to mid-scale by using P-13.

g) Repeat steps e and f until a satisfactory balance and level adjustment is achieved. An accurate adjustment is not necessary since the trace can be positioned finally by using either the OUTPUT ADJUST control (for SENSITIVITY position 1), the BALANCE control (for SENSITIVITY positions 2, 3, 4, and 5), or the oscilloscope Vertical Position control.

4. Sensitivity

The nominal sensitivity of the amplifier for the various positions of the SENSITIVITY switch is as follows:

Position	*Sensitivity
1	100 millivolts
2	10 millivolts
3	1 millivolt
4	250 microvolts
5	50 microvolts

*This is the input signal required to give a 1 volt output at the OSCILLOSCOPE terminal and 4.8 volts across the METER terminals.

5. **Response**

The frequency response obtainable from the OSCILLOSCOPE terminals is limited by the RESPONSE switch. The frequency at which the response is 70% of that at d-c is as follows:

Position	70% Response Frequency
1	500 cps.
2	1000 cps.
3	5000 cps.
4	10000 cps.
5	maximum available (40 to 60 kc)

APPENDIX II

The wiring diagrams of Figs. 25 and 26 give the interconnections between the various chassis. Wiring Diagram I (Fig. 25) shows the Power Supply Chassis and Regulator Chassis and the cables connected between them. Those components included within the dotted line portion of the Regulator Chassis are mounted on the vibration mounted sub-chassis.

The Wiring Diagram II (Fig. 26) gives the interconnections between the Output Chassis and the Amplifier Chassis. For each chassis the panel controls are isolated, and the components are drawn in their approximate location on the chassis as seen from the bottom. Only one amplifier and output channel are shown; the other is a mirror image of the one illustrated. The components mounted on the vibration mounted sub-chassis are those in the dotted line section of the Amplifier Chassis.

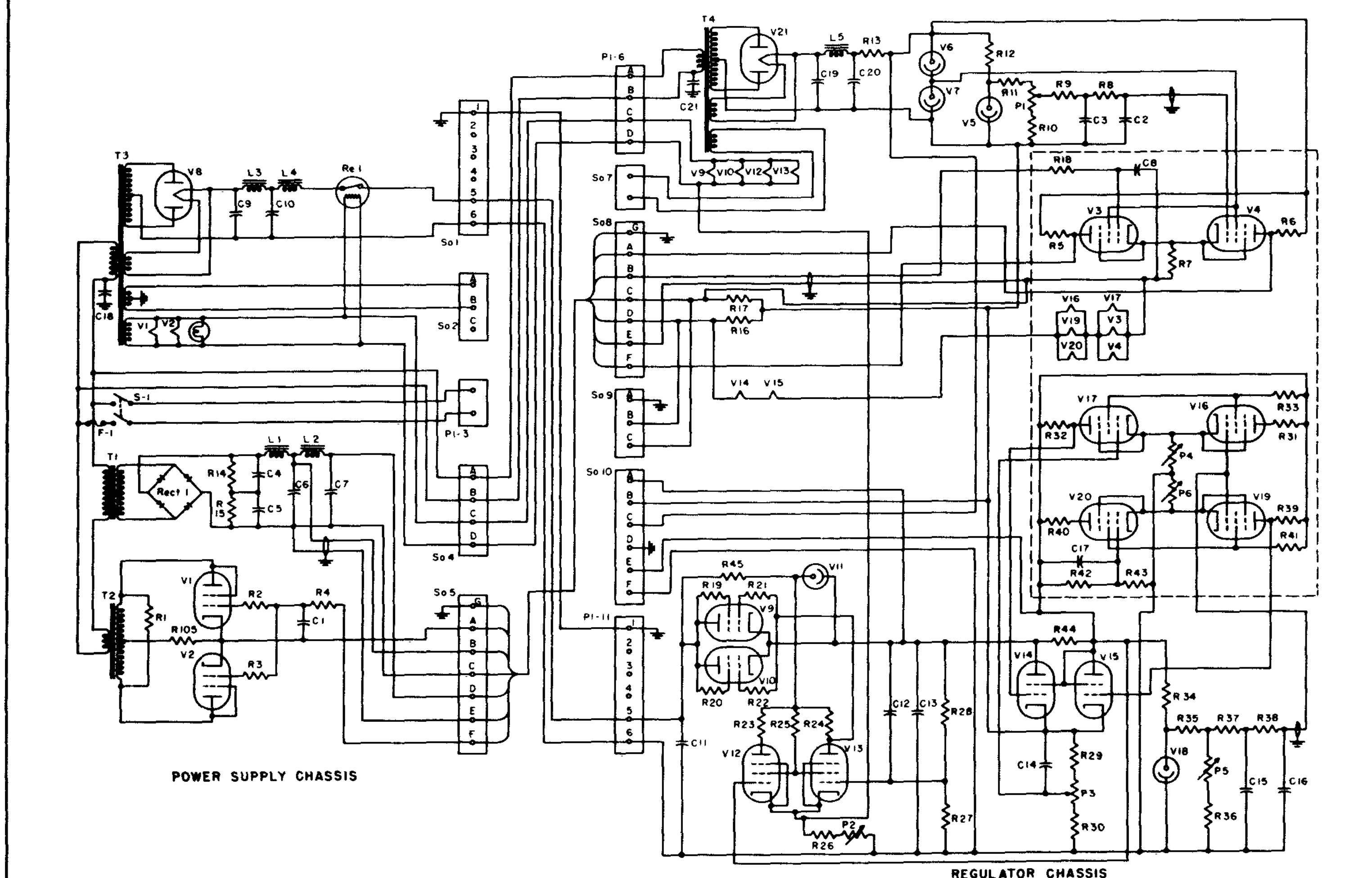

FIG-25 WIRING DIAGRAM I

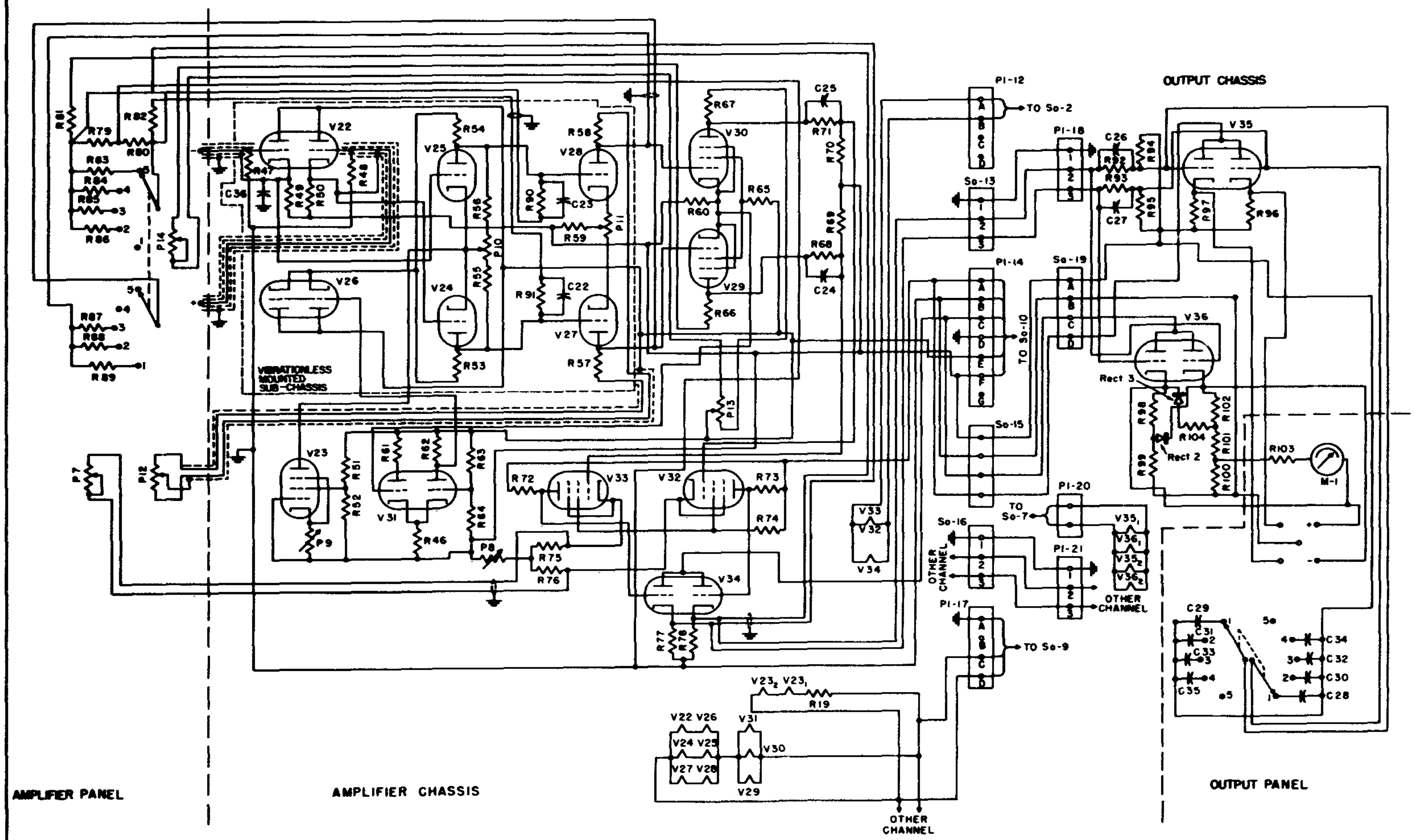

FIG-26 WIRING DIAGRAM II